ELECTRICAL COMMUNICATION

1960
VOLUME 36
NUMBER 4

THE TECHNICAL JOURNAL OF ITT

INTERNATIONAL TELEPHONE AND TELEGRAPH CORPORATION

ELECTRICAL COMMUNICATION

Technical Journal Published Quarterly by
INTERNATIONAL TELEPHONE and TELEGRAPH CORPORATION
67 Broad Street, New York 4, New York

President: H. S. Geneen

Secretary: E. J. Whalen

Subscription: $2.00 per year; 50¢ per copy

EDITOR
H. P. Westman

ASSISTANT EDITOR
J. E. Schlaikjer

EDITORIAL BOARD
H. G. Busignies
R. S. Caruthers
G. Chevigny
A. G. Clavier
E. M. Deloraine
F. R. Furth
G. Goudet
B. C. Holding
J. Kruithof
W. P. Maginnis
A. W. Montgomery
I. N. Palley
E. D. Phinney
G. Rabuteau
P. C. Sandretto
T. R. Scott
C. E. Strong
F. R. Thomas
H. B. Wood

Copyright © 1960 by
INTERNATIONAL TELEPHONE and TELEGRAPH CORPORATION

ELECTRICAL COMMUNICATION

The Technical Journal of
INTERNATIONAL TELEPHONE AND TELEGRAPH CORPORATION
AND ASSOCIATE COMPANIES

CONTENTS

VOLUME 36 **1960** **NUMBER 4**

Modern Power-Tube Techniques 229

K24A Syncroplex Telephone Carrier System *by B. G. Coetsee, G. L. Curtis, and J. W. Halina* 230

High-Speed Transmission of Numerical Data Over Telephone Channels *by Anatole Girinsky and Paul Roussel* 248

Some Performance Parameters of Silicon Junction Power Rectifiers *by D. R. Coleman* 263

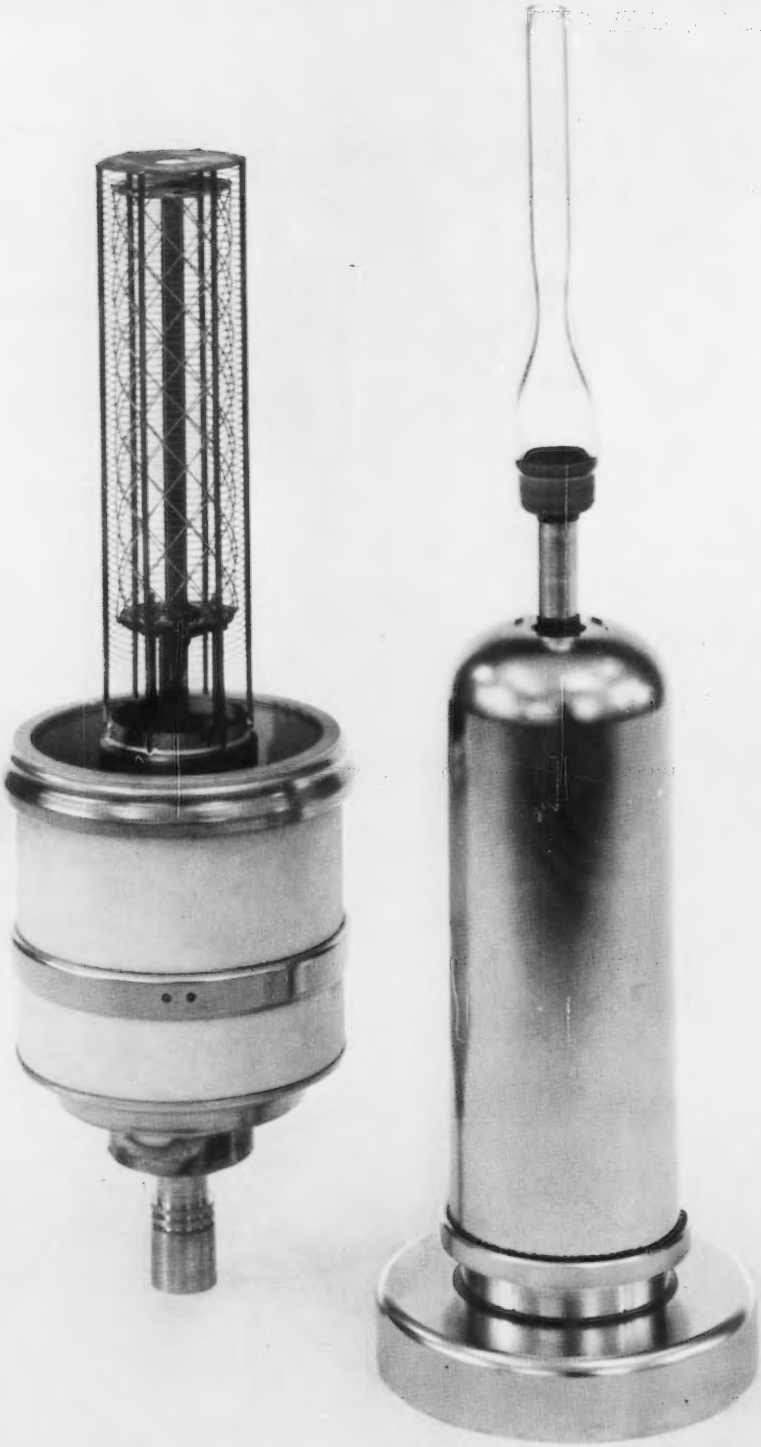
Phototubes Capable of High Current Output *by Sanford F. Essig* 272

AN/MRC-55 General-Purpose Mobile High-Frequency Communication Equipment *by M. C. Dettman* 276

United States Patents Issued to International Telephone and Telegraph System; November 1959–January 1960 285

In Memoriam—Sir Francis J. E. Brake 290

Contributors to This Issue 291



Modern Power-Tube Techniques

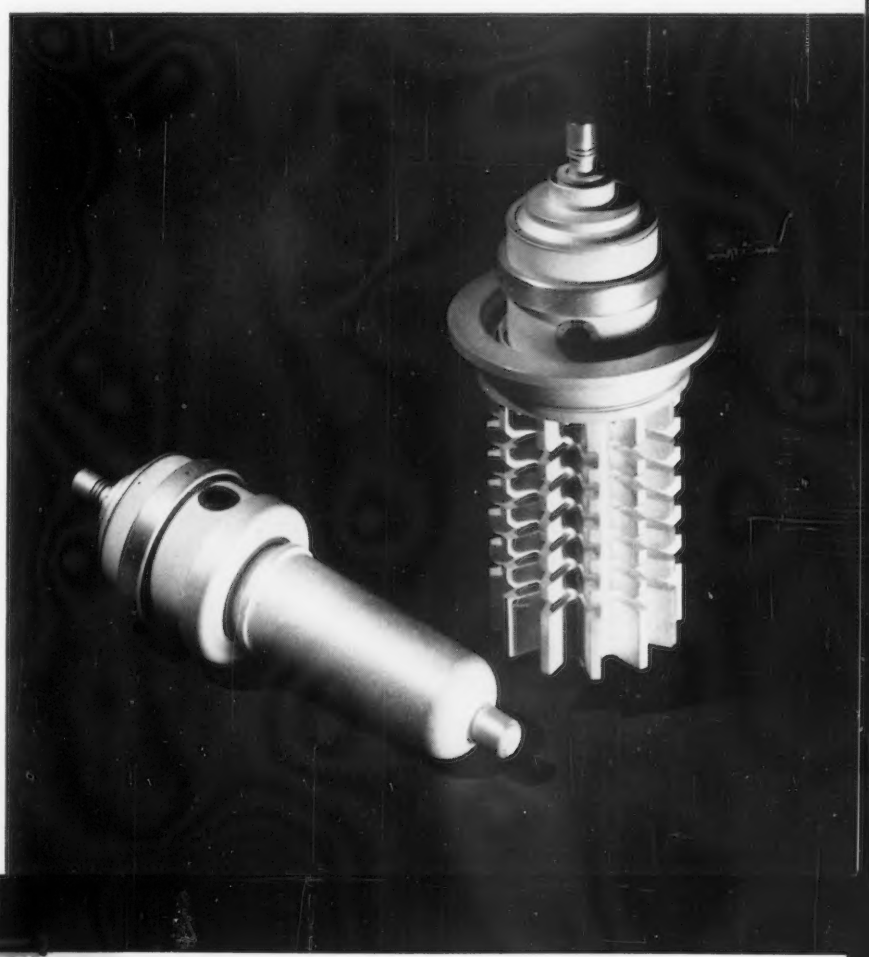
THE *D-1008* is a general-purpose metal and ceramic triode built by the ITT Federal Division at Clifton, New Jersey, using the latest techniques in power-tube manufacture. It may be used as a class-*C* amplifier and oscillator under continuous-wave or pulsed conditions. Its coaxial design permits use in cavities. Maximum ratings of 60 kilowatts plate input and 10 kilovolts of plate voltage apply up to 50 megacycles per second. It is also capable of delivering a peak pulse power of two megawatts as a hard-tube modulator.

Superior performance from a tube is the result of superior individual characteristics. The high-alumina ceramic envelope improves the potentialities of the tube by providing a mechanically strong high-electrical-resistance vacuum-tight barrier. Because of its low-loss characteristics in radio-frequency fields and its excellent high-temperature stability, dielectric heating of the envelope, a prime cause of catastrophic failure of glass-envelope tubes in high-frequency application, is eliminated.

To minimize the heating effects of large circulating currents at the anode and grid seals during high-frequency operation, copper was selected as the most-desirable structural material. The external metal surfaces are also silver plated to prevent oxidation. As shown in the frontispiece, the cathode consists of two multifilar helixes of thoriated tungsten wire angled with respect to each other and welded to form a mesh of cylindrical form. This design provides ample emission at conservative filament temperatures. As the emission life of the cathode depends on its operating temperature, this is essential to long tube life. In addition, the mesh construction is mechanically strong and resists distortion. Correct interelectrode spacing is preserved as a result and tube characteristics remain unchanged throughout tube life.

The grid is of swaged construction; that is, the grid supports are notched, the grid turns are laid in the notches, and are secured by offsetting the metal at the sides of the notch. This method eliminates possibilities of burnt welds, cold welds, and embrittlement of the grid turns associated with grids secured by spot welding. The grid rods are arc welded to the support cone that in turn is brazed to the copper grid ring; thus a strong, low-resistance thermal and electrical path is assured.

Two types of anode are used as shown in the photograph on this page. The tube with the conventional anode is cooled by circulated water and is capable of dissipating 20 kilowatts in continuous commercial service. Tube type *D-1028* is identical to the *D-1008* except for the anode, which has been designed for evaporative cooling and is rated at 30 kilowatts. The over-all height of the tube is $13\frac{3}{4}$ inches (35 centimeters).



K24A Syncroplex Telephone Carrier System*

By B. G. COETSEE, G. L. CURTIS, and J. W. HALINA

ITT Laboratories, a division of International Telephone and Telegraph Corporation; Palo Alto, California

1. *Background and Description*

A MOST-SIGNIFICANT turning point in the history of telephone transmission occurred shortly before 1920 with the introduction of carrier equipment as a means of multiplying the capacity of open-wire line facilities. The 20-year period between 1918 and 1938 witnessed the firm establishment of single-sideband techniques in telephone transmission through the *C* carrier and the development of the 12-channel *J* and *K* systems for open wire and cable, respectively. This was a period of rapid expansion of carrier applications to long-haul toll circuits and, by and large, a consolidation of carrier design for this type of service. If the *L* system is included, which is essentially an expansion of the *K*, the equipment supporting the majority of long-haul toll transmissions is still of the *J*, *K*, or *L* type.

A new departure in carrier equipment development occurred around 1950 with the introduction of carrier to short-haul toll circuits, as well as to tributary, trunk, and long interexchange lines. New techniques in design and manufacture produced the *N*, *O*, and *ON* equipments, a class of systems that made economically feasible the application of carrier to circuits between 20 and 200 miles (32 and 320 kilometers) long. With the advent of these carrier systems, the number of carrier channel miles in the nation advanced to a new order of magnitude.

As this application of carrier gained momentum, the attention of telephone communication engineers turned to an area of the telephone transmission plant that had heretofore defied the economic application of carrier techniques on any significant scale; that is, the exchange trunk circuit where more than 50 percent of the circuits fall in the 5-to-20-mile (8-to-32-kilometer) range. If a carrier equipment could be produced to serve this class of circuit, still another order of magnitude of carrier application to telephone transmission could be achieved, with its attendant gain in service to the subscriber.

After a study of design possibilities and available techniques, a decision was made at ITT Laboratories in mid-1957 to proceed with the development of an exchange carrier system. This culminated in the field trials of the *K24A* syncroplex system toward the end of 1958. Because of the tremendous demand for circuits, an objective was set to compress the usual 4-to-5-year cycle of development and testing of a basically new carrier system into 1½ to 2 years.

1.1 Characteristics of Trunk and Exchange Transmission Facility

A most-important aspect of the design and development problem in the telephone transmission field is that of a large existing facility that it is desirable to improve, but which it is virtually impossible to redesign in any radical way. This, then, means the development of equipment to work with existing facilities that were frequently designed and installed at great cost for a different use than that now intended. If one were starting over again, more-suitable facilities could be designed, but this is small comfort.

The exchange trunk facilities are essentially low-frequency mediums. For voice transmission, these lines are equalized for response in the 0-to-4-kilocycle-per-second frequency range by the addition of loading coils. Repeaters of the hybrid or negative-impedance type are employed to provide gain where necessary. In addition to voice, the physical line is, of course, capable of transmitting tone signals and direct-current pulses and all possible states of the loop battery; that is, battery on, off, normal, or reversed. A carrier system to replace this service economically would have to be capable of using the facility to about 0.5 megacycle per second, accommodate

* Presented before the Fall General Meeting of the American Institute of Electrical Engineers, Chicago, Illinois; October 11–16, 1959. Reprinted from *Communications and Electronics*, number 46, pages 1044–1054; January, 1960.

a relatively complicated signaling transmission requirement, provide a very-high degree of quality and reliability, be simple to install and maintain, and be one half to one quarter as costly as previously applied carrier systems—a truly imposing challenge.

The electrical characteristics of typical exchange lines are depicted[1] in Figure 1. The attenuation characteristics are shown in Figure 1A and the thermal properties in Figure 1B. Three other characteristics of great importance are near-end crosstalk, far-end crosstalk, and impulse noise power. These are parameters that, due to the nature of the medium, can only be described statistically. On the basis of very-limited tests, the range in which these parameters fall is shown in Table 1.

[1] Reproduced by permission from a paper by E. F. Vaege, "Cable Transmission Characteristics for the *N* Carrier System," *Bell Laboratories Record*, volume 30, pages 333–336; August, 1952. Extrapolations for frequencies beyond 250 kilocycles per second are the responsibility of the authors.

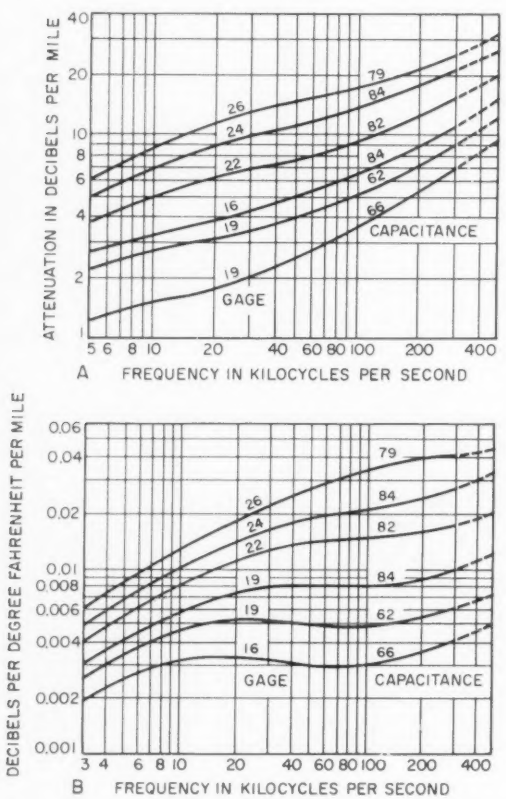


Figure 1—Typical exchange lines. *A* = attenuation characteristics at 55 degrees fahrenheit and *B* = thermal characteristics. Capacitance is given in nanofarads per mile.

TABLE 1

CABLE CROSSTALK CHARACTERISTICS

Parameter, at 500 Kilocycles per Second	Percentage of Pairs in a Cable With Characteristics Worse Than Value Indicated			
	100	50	25	10
Near-end Crosstalk Coupling Loss in Decibels	95	82	74	68
Far-end Crosstalk Coupling Loss in Decibels per Mile	85	63	56	52
Root-Mean-Square Impulse Noise in Decibels Relative to 1 Milliwatt	—	−100	—	—

From an assessment of these data, it follows that an exchange carrier system must have the following principal characteristics:

(A) Economical application on line lengths between 5 and 20 miles (8 and 32 kilometers), or attenuations of the order of 200 decibels at 500 kilocycles per second.

(B) Repeaters with slope equalization in the low-frequency-to-500-kilocycle-per-second range of from a few decibels to a maximum of approximately 16 decibels.

(C) Adequate system regulation for 15-to-18-decibel variation in 175-to-200-decibel lines.

(D) Transmission disturbances of a maximum of 20 decibels adjusted (dba) for all types of noise.

The fourth requirement implies that for a 15-mile (24-kilometer) system, the minimum signal level would have to be more than 37 decibels below 1 milliwatt (dbm) loss from 2-wire input to lowest operating level (lower, to the extent that the modulation system chosen would have some improvement factor). In addition, the system would have to provide for 15 to 25 decibels improvement on far-end crosstalk for about 25 percent of the pairs within a cable and 10 to 20 decibels of improvement on near-end crosstalk, again for about 25 percent of the pairs, depending on level differentials between the maximum transmitting power and the minimum receiving power.

In consideration of power and space economies, as well as design obsolescence, it was a foregone conclusion that design would be based on transistors rather than vacuum tubes.

1.2 Choice of Modulation System

One of the obvious answers to many of the transmission problems would have been the choice of a modulation system capable of trading noise for bandwidth, such as pulse code. However, it was apparent that a pulse-code-modulation design was more expensive than a frequency-division-multiplex approach and was less predictable of technical success. On the other hand, consideration of the near-end crosstalk and available power-handling capacities of high-frequency transistors indicated that a modulation scheme that was reasonably conservative of spectrum and reasonably efficient with regard to multichannel loading would be desirable. A modulation system not requiring a wide bandwidth per channel would permit an interleaved modulation spectrum for near-end-crosstalk immunity without using more than 500 kilocycles per second of spectrum for the 24 channels. If, in addition, it did not require a power-consuming carrier, it would be possible to make efficient use of readily available transistors that offer a peak power-handling capacity of the order of +13 decibels relative to 1 milliwatt.

At about this time, a development at ITT Laboratories produced an economical means of detecting double-sideband suppressed-carrier amplitude modulation. Although it was recognized for many years that this type of modulation could offer a close approach to the performance characteristics of single-sideband suppressed-

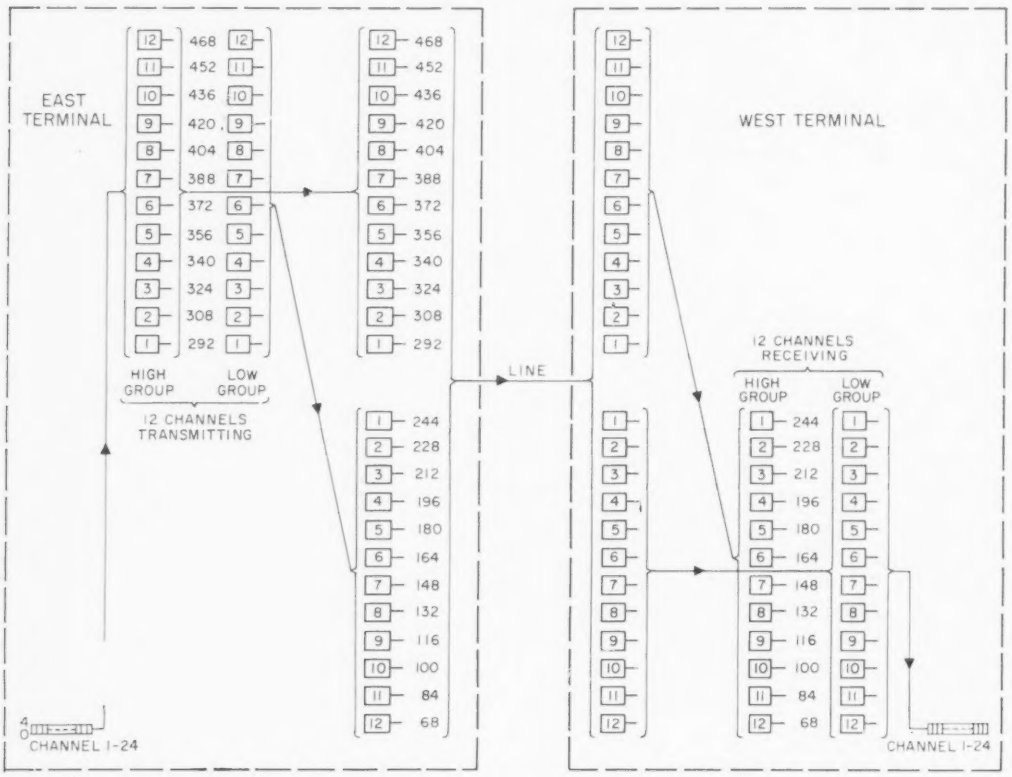


Figure 2—*K24A* modulation plan for east-to-west transmission. All frequencies for west-to-east transmission are 8-kilocycles-per-second higher. The boxes indicate the channels and the corresponding frequencies are given in kilocycles per second.

carrier amplitude modulation in the matters of band conservation and loading factors while avoiding the high channel filter costs of single sideband, its use had been hampered by the lack of a means of detection sufficiently economical to warrant its choice over single sideband in consequence of its use of twice the bandwidth

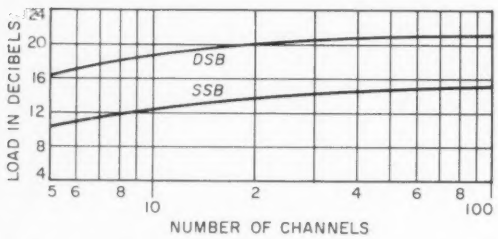


Figure 3—Comparison of load factors. *DSB* = double-sideband suppressed-carrier transmission with peak limiting. *SSB* = single-sideband suppressed-carrier transmission with peak limiting. The ordinate is the root-mean-square test tone load capacity in decibels above channel sideband zero-level reference.

and its fourfold greater loading factor. In view of the laboratory development of a detector, it appeared that a modulation system was available to satisfy all but one of the combined requirements of an exchange system—far-end crosstalk on about 40 or 50 percent of arbitrarily selected pairs. It was decided that the system would be so designed that a compandor or some other type of noise-reducing circuit could be introduced by the simple substitution of a plug-in panel. In this manner, a system was available that, for 50 to 60 percent of the arbitrarily chosen pairs, would not require the expense of noise-suppression equipment, while, at the same time, the infrequent incidence of pairs with higher crosstalk would permit simple treatment by the substitution of the voice-frequency panel by an alternative one containing circuits furnishing the necessary noise suppression. A Sonad† (speech-operated noise-attenuation device), providing an additional 14–18 decibels of noise suppression during idle-circuit periods, was developed at ITT Laboratories and is undergoing extensive tests to determine its application possibilities. For larger improvements, 20–23 decibels, a syllabic compandor is available.

† Trademark.

The modulation plan and frequency assignments are shown in Figure 2, from which it is seen that the system is based on an 8-kilocycle-per-second spectrum occupancy per channel, with a spacing of 16 kilocycles per second between channels, allowing 24 channels between 60 and 480 kilocycles per second. The 16-kilocycle-per-second spacing permits interleaving of reverse direction of transmission and thus provides for immunity to near-end crosstalk. This wide spacing also eases channel separation filter requirements.

The loading characteristic is shown in Figure 3 in comparison with single-sideband suppressed carrier. With a 20-decibel loading factor for 24 channels and a transistor-repeater overload capacity of +13 decibels relative to 1 milliwatt, it was possible to transmit at a level of −12 decibels per channel and accomodate a better-than-25-decibel range without dropping below the established minimum signal level of −37 decibels.

Regulation was incorporated in the repeaters to provide a stability of ±1 decibel for an 8-decibel line-level variation and in the channel terminals a stability of ±0.5 decibel for an 8-decibel variation, resulting in a level stability of ±0.5 decibel for a 175-decibel system. Regulation is not affected by the addition of Sonad circuits. In the case of channels with compandors, the stability is 1.0 to 1.5 decibels under the same line conditions.

Equipment linearity keeps all types of distortion below 20 decibels adjusted for an 8-repeater system.

The system block diagram is shown in Figure 4.

1.3 Signaling

An exchange carrier system must provide a choice of signaling facilities not previously provided as an integral part of carrier systems. Two types of signaling are of principal interest. In their simplest terms they require the transmission of the following information:

(A) *Loop signaling*—This signaling option requires the transmission of closed-loop–open-loop conditions from an originating to a terminating office at a rate that may be as high as 50 alternations per second. In the reverse direction, the transmission is that of the condition of battery, normal or reversed, at a somewhat slower rate.

(B) *Revertive signaling*—In this signaling option, the originating-to-terminating-office transmission consists of open-loop–closed-loop conditions at up to an equivalent of 60 alternations per second and, in the reverse direction, all states of loop battery; normal, off, reversed.

Moreover, it provides a means for channel regulation.

The resulting signaling facility is, in fact, an excellent telemetering channel, capable of handling up to 80 bits per second with a pulse distortion of less than 5 percent.

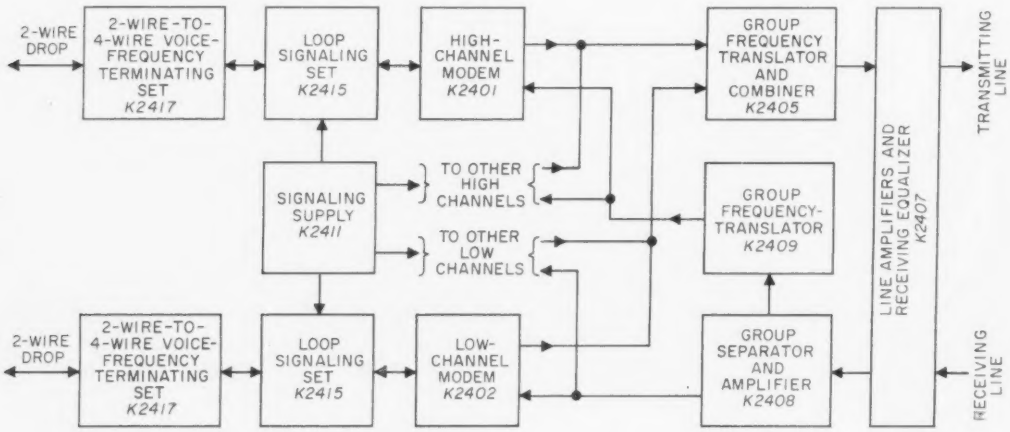


Figure 4—*K24.1* terminal. More-detailed descriptions of the functions of each module can be obtained by comparison with the type numbers given in Table 2.

The familiar *E* and *M* signaling facilities normally offered by carrier systems must be augmented by a 4-wire-to-2-wire, or a 4-wire-to-3-wire converter to provide loop signaling. The third wire in the 3-wire option furnishes supervisory information normally provided by the outgoing trunk circuit and, when used, eliminates the need for this appliqué panel. In the case of the revertive system, a facility capable of transmitting one bit more information from the terminating office to the originating office is necessary, a requirement of one bit more than *E*-and-*M* or loop-signaling systems are capable of handling. Despite the fact, therefore, that a double-sideband suppressed-carrier amplitude-modulation system offers an economical one-bit transmission system with the availability of the carrier for control-information transmission purposes, the need for providing revertive signaling led to the choice of frequency-shift signaling, which is capable of transmitting more than one bit of information as it offers the alternatives of sending either one of the two frequencies, both, or none.

1.4 Mechanical Aspects of *K24.1* System

In a search for the simplest and most-flexible mechanical arrangement, a computer-type construction was evolved, as shown in Figure 5. The system features printed-circuit wiring, "banjo"-type rack wiring and plug-in "bookshelf" construction. As indicated earlier, each channel is subdivided to permit easy assembly of a system for any application; signaling cards can be readily substituted as required without necessitating realignment of the remainder of the system. Similarly, the voice-frequency terminating unit can be readily changed to provide 2-wire, 4-wire, compandor, Sonad facilities, et cetera.

1.5 Summary of System Parameters

Following is a summary of the *K24.1* system parameters:

Number of channels: 24

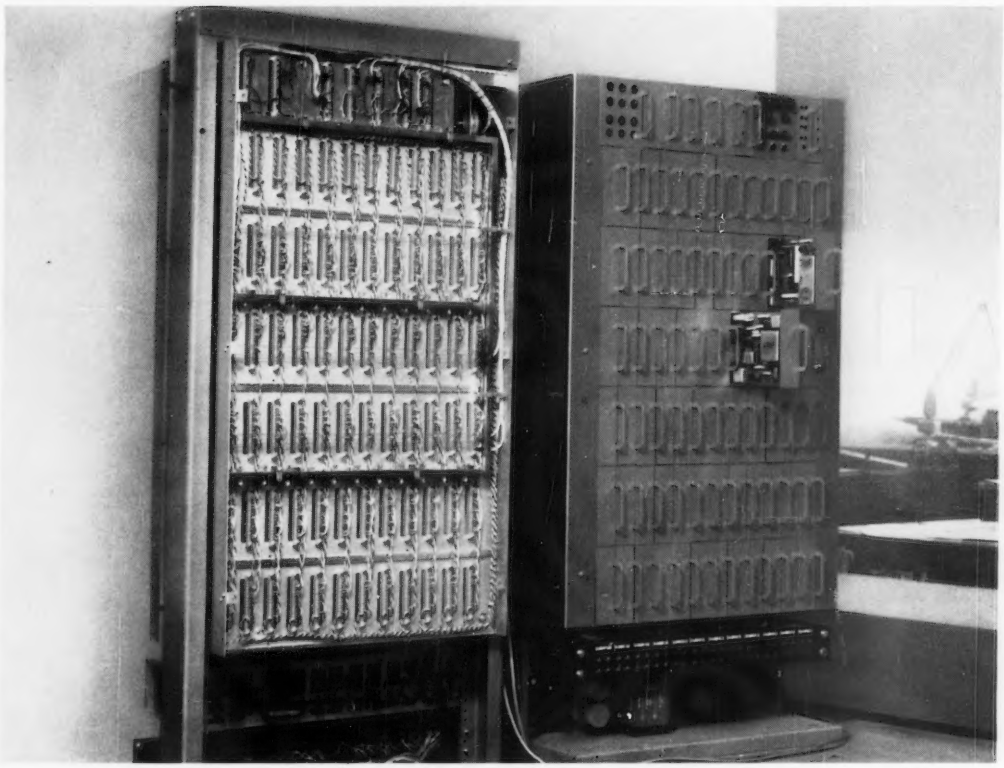


Figure 5—Prototype of *K24A* terminal equipment.

Frequency spectrum: 60 to 480 kilocycles per second

Transmitting level per channel sideband: −12 decibels relative to 1 milliwatt

Receiving level per channel sideband: −37 decibels relative to 1 milliwatt

24-channel loading factor: 20 decibels

Maximum repeater power capacity: +13 decibels relative to 1 milliwatt

Span: attenuation, 25 decibels maximum; slope, 18 decibels maximum

System regulation: ±0.5 decibel

Channel equivalent: 0 or −3 decibels relative to 1 milliwatt

Terminal noise: 20 decibels adjusted at zero level (maximum all types)

System noise: 25 decibels adjusted maximum for 175 decibels

Signaling: *E* and *M*, loop, or revertive

Compandors: optional

Sonad: optional

Size: 31.5 inches (80 centimeters) of rack space per 24-channel terminal

Power consumption: approximately 100 watts at 48 volts per 24-channel terminal

Repeater power consumption: 56 milliamperes at 30 volts; optional, regulated or nonregulated

Alarms: power failure, common equipment failure, and system failure

2. Equipment Considerations

The *K24A* syncroplex carrier telephone system is a 24-channel multiplex equipment based on

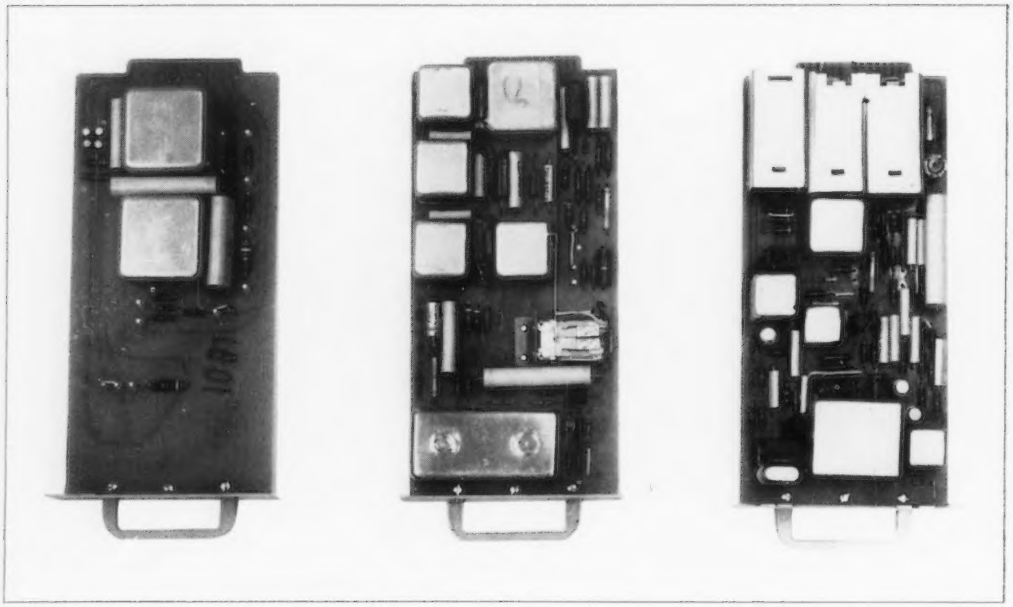


Figure 6—*K24.1* channel equipment. From left to right voice-frequency terminating module, the loop signaling module for the originating office, and the channel modem.

double-sideband suppressed-carrier modulation, in which a base group of 12-channel modulators (transmitters) operates at carrier frequencies between 292 and 468 kilocycles per second and a corresponding base group of 12-channel demodulators (receivers) operates between 68 and 244 kilocycles per second. Each channel occupies 8 kilocycles per second of spectrum; thus adjacent channel spacing can be as close as 8 kilocycles per second. In the case of the *K24.1* system, which is arranged for exchange-cable applications, the spacing is 16 kilocycles per second to permit interleaving of the two directions of transmission and to prevent crosstalk as explained in section 1.

To complete the transmission path between the 12-channel modulator group and the 12-channel demodulator group, it is thus necessary to translate (group-modulate or demodulate) once.

For consideration of the equipment design, the *K24.1* system can be divided into four parts:

(A) *Channel terminal equipment*, shown in Figure 6, consists of the modem unit that derives the channel signal from the multichannel spectrum, the signaling unit that permits the transmission of various types of switching information, and the terminating unit providing 4-wire-to-2-wire conversion for the signaling circuit, as well as other functions such as voice-frequency companding, signal limiting, et cetera.

A great variety of arrangements of channel terminals is possible through the appropriate choice of signaling and voice-frequency terminating units, yielding excellent flexibility for many system applications.

(B) *Group equipment*, shown in Figure 7, comprises the 12-channel group frequency translating equipment and a 24-channel line amplifier.

(C) *Auxiliary equipment*, including the terminal voltage regulator, the fuse-and-alarm unit, and the 24-channel signaling-tone supply.

(D) *Line equipment*. For operation over exchange cables, there are provided repeaters of both regulating and nonregulating types, equipped with span pads and equalizers.

TABLE 2

FUNCTIONAL DESCRIPTION OF *K24A* UNITS

Group	Unit	Option	Function	Type
(A) Channel Terminal Equipment	Channel Modem	Master	Channel transmitter and receiver including crystal-controlled carrier supply for two transmitters	*K2401*
		Slave	Like *K2401*, except without carrier oscillator	*K2402*
	Signaling Unit	*E* and *M*	Frequency-shift transmitter and receiver discriminator, including signaling-tone pick-off filter, automatic-gain-control rectifier, and relay	*K2419*
		Loop, 2-Wire	Like *K2419* except arranged to operate from and into *a-b* leads of voice-frequency terminating set. Provided in originating and terminating office versions	*K2415–6*
		Loop, 3-Wire	Like *K2415–6* except control lead added at originating office to eliminate need for outgoing pulse repeaters in step-by-step offices	*K2425–6*
		Revertive	Like *K2415–6* except equipped to provide channel for revertive pulse transmission	*K2403–4*
	Voice-Frequency Terminating Unit	2-to-4-Wire Terminating Set	2-to-4-wire voice-frequency hybrid, *a-b* lead connections, voice-frequency limiter	*K2417*
		2-to-4-Wire Terminating Set and Sonad	Like *K2417* except Sonad added	*K2418*
		2-to-4-Wire Terminating Set and Compandor	Like *K2417* except compandor added	*K2428*
(B) Terminal Common Equipment	Group Modulator		Translates 12 high-group transmit channels to low-group spectrum, combines both groups, generates pilot tone	*K2405*
	Group Receiving Amplifier		Separates high group from low group, amplifies low group for channel detection	*K2408*
	Group Demodulator		Translates 12 high-group received channels to low-group spectrum and amplifies the group for channel detection	*K2409*
	Line Amplifier		Amplifies channel-to-line level for transmission, provides span adjustment, equalization and amplification for received signals. Includes facilities for connecting repeater power	*K2407*
(C) Terminal Auxiliary Equipment	Signal Supply		Low-impedance power source of 3.7-kilocycle-per-second and 3.9-kilocycle-per-second tones for 24 signaling units	*K2411*
	Fuse and Alarm Panel		Contains fuses, power-supply failure alarm, pilot failure alarm, common equipment, and channel failure alarm	*K2406*
	Power Supply Regulator		Converts −48-volt direct-current supply to −30-volt regulated supply	*K2412*
(D) Repeater Equipment	Repeater	Regulator Repeater	25-decibel span amplifier including span pads and equalizers	*K2414*
		Nonregulator Repeater	Like *K2414* but without regulating circuit	*K2413*
(E) Terminal Shelf			Accommodates the 79 panels composing a 24-channel terminal, terminal strip, and repeater supply current control	*K2410*
(F) Test, Line-up, and Maintenance Equipment	Alignment Set		Facilities for testing repeaters and line sections for span length adjustment, equalization, and trouble shooting	*K2480*
	Unit-Extension Test Sets		Facilities for extending units out of shelf and making test measurements	*K2481–3*

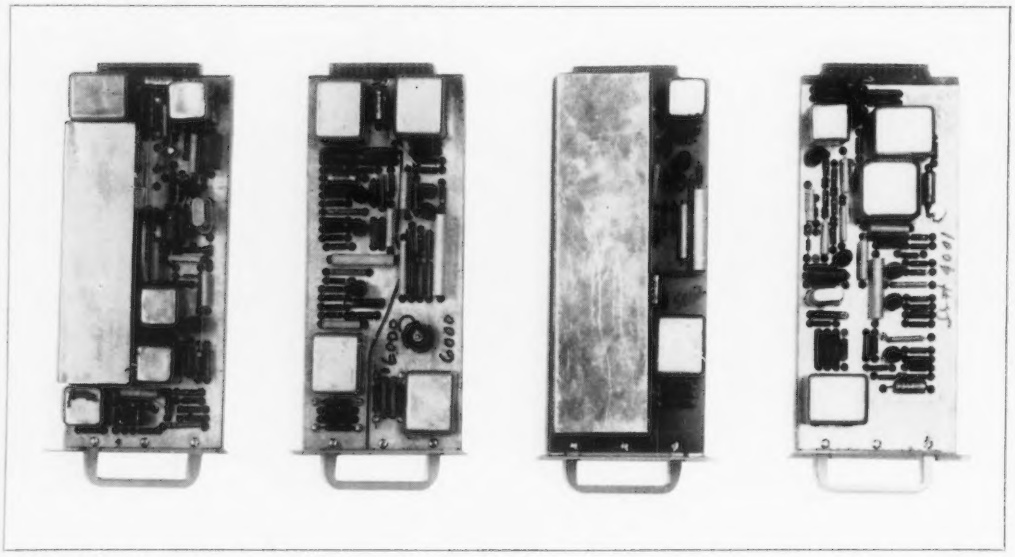


Figure 7—*K24.1* group equipment. From left to right are the group modulator, line amplifier, receiving amplifier, and a group demodulator.

For ease of reference, the entire complement of units and their functional descriptions are summarized in Table 2.

2.1 Circuits

The *K24.1* system uses transistors throughout. This represents one of the first large-scale applications of transistors to carrier telephony of up to 0.5 megacycle per second of carrier spectrum. Many stringent performance requirements, which are probably more peculiar to multichannel telephony than to other areas of communication, have to be met. Accordingly, several basic circuits were developed and used

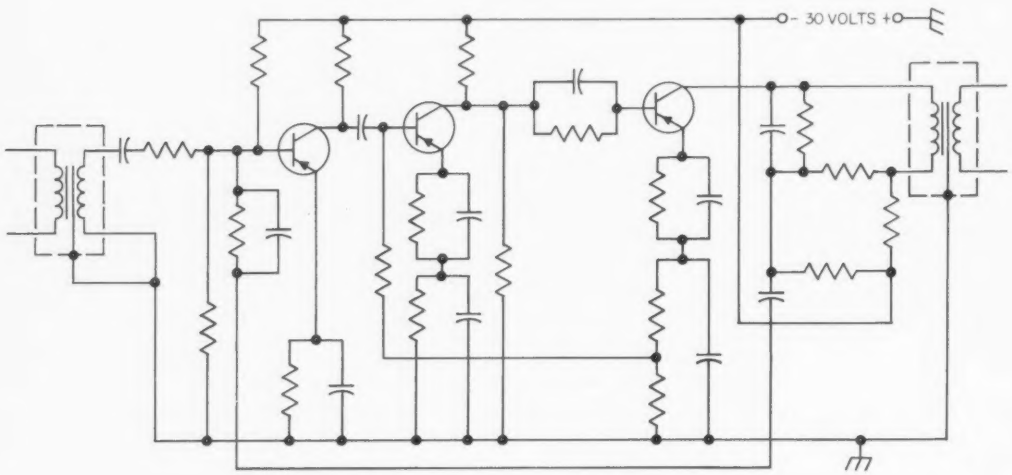


Figure 8—Simplified schematic diagram of the 32-decibel amplifier.

in a variety of functional units. The principle of these units will be described now and referred to later in the description of units where they occur.

2.1.1 32-Decibel Amplifier[2-4]

One circuit is a 3-stage 60–480-kilocycle-per-second wideband amplifier with an average of 25 decibels of negative feedback in the range of application. An interesting aspect of this unit is that the transistors used in it exhibit a mean beta cutoff of 250 kilocycles per second, whereas the amount of feedback necessary to achieve linearity required a controlled feedback characteristic to beyond 4 megacycles per second.

A simplified schematic of this amplifier is shown in Figure 8. It consists of three common-emitter stages in cascade. The design was accomplished by successive *RC* compensation of each of the three stages to maintain a 30-degree phase margin to beyond the point of unity gain in the feedback loop. The output impedance is bridge stabilized, and the input impedance is stabilized by passive resistance. The result is an over-all impedance stability of better than 5 percent across the operating bandwidth.

The feedback and nonfeedback gain characteristics of the amplifier are shown in Figure 9A, the load characteristic in Figure 9B, and the linearity characteristic in Figure 9C.

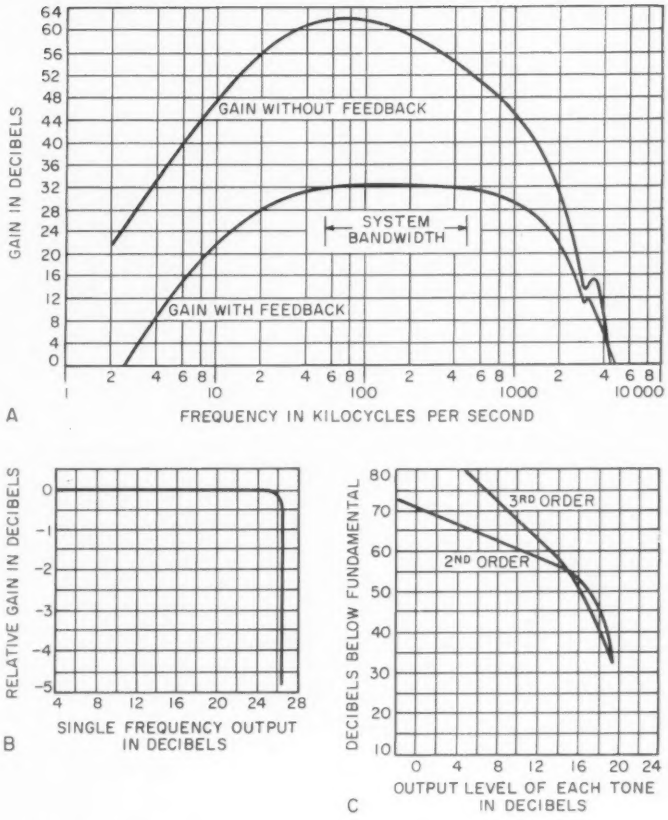


Figure 9—Characteristics of the 32-decibel amplifier: *A* = frequency response. *B* = loading characteristics. *C* = intermodulation products. Output levels in *B* and *C* are in reference to a test tone output of 12 decibels below 1 milliwatt.

The amplifier is the basic circuit in the regulating and nonregulating repeaters, the transmitting section of the line amplifier, and the group demodulator.

2.1.2 15/19-Decibel Amplifier

Except for gain, the characteristics of this amplifier; linearity, feedback, and load-handling capacity; are the same as those of the 32-decibel amplifier. The lower gain required permits the use of two stages, a common-base and a common-emitter, in cascade. The amplifier is impedance stabilized with bridge feedback at both output and input.

[2] F. D. Waldhauer, "Wideband Feedback Amplifiers," *IRE Transactions on Circuit Theory*, volume PGCT-4, pages 178–189; September, 1957.

[3] R. P. Abraham, "Wideband Transistor Feedback Amplifiers, *WESCON Convention Record*, part 2, pages 10–19; 1957.

[4] F. H. Blecher, "Design Principles for Single-Loop Transistor Feedback Amplifiers," *IRE Transactions on Circuit Theory*, volume PGCT-4, pages 145–156; September, 1957.

2.1.3 *Crystal-Controlled Carrier-Supply Oscillator*

The requirement for stable medium-power constant-voltage sources of carrier supply exists in many units of the system; channel modulators, group modulators and demodulators, pilot generators, et cetera.

The circuit chosen[8] was that of a modified Colpitts (commonly referred to as the Clapp or Gouriet) oscillator, direct-current coupled to a bootstrap amplifier. A simplified circuit is shown in Figure 10.

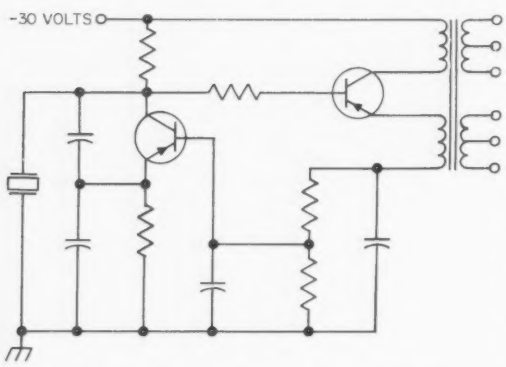


Figure 10—Simplified schematic of carrier supply.

The use of a bootstrap buffer provides a device with a very-high-impedance load on the crystal oscillator, a very-low output impedance that permits driving multiple loads with low coupling, and good power gain and overload characteristics. The resultant carrier supply is capable of 20-to-24 milliwatts of output power from a source impedance of a few ohms.

2.1.4 *LC Oscillator*

Due to the cost or size of low-frequency crystals, or the inflexibility of crystal oscillators, a stable *LC* oscillator circuit is desired in such units as the signaling-tone supply and the local-oscillator source in the channel modem. For these applications, a tuned collector–emitter feedback circuit was designed. The oscillator operates at frequencies of 3.7 to 252 kilocycles per second and, without the aid of a buffer, provides over 25 milliwatts of power with a frequency stability of better than 0.1 percent over a temperature range of 0–60 degrees centigrade.

2.1.5 *Filters*

Considerations of economy in filters are all-important in carrier equipment design, particularly at the channel-modem level where filters are numerous. The structure used in the *K24.1* system is that of an intermediate-frequency-amplifier-type filter in which coil construction with adjustable leakage coupling are used to achieve the elements of a mesh. Figure 11 shows the transmitting filter, and Figure 12 typical insertion-loss curves of the filters.

The simplicity of the receiving filters results in low selectivity, but this can be tolerated because both adjacent channels are demodulated out-of-band with synchronous detection; in single sideband with normal channel spacing, one of the adjacent channels will be demodulated in-band. With emitted-carrier double-sideband and envelope detection, all channels are demodulated in-band.

2.2 Units of *K24.1* System

The functions of the various units of a *K24.1* terminal, briefly described in Table 2, are further

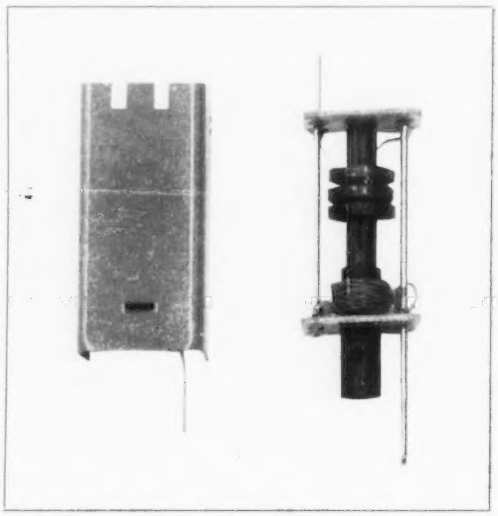
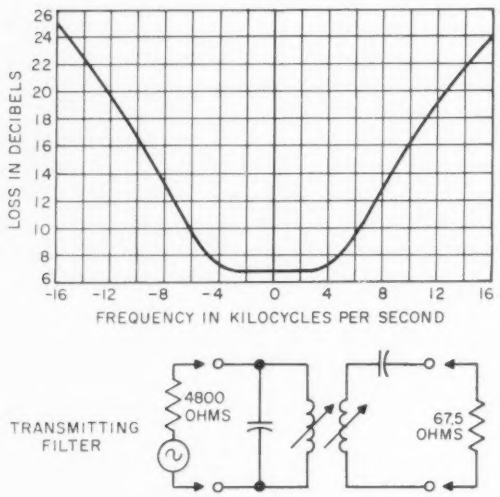


Figure 11—*K24.1* channel transmitting filter.

[8] E. Keonjian, "Stable Transistor Oscillator," *IRE Transactions in Circuit Theory*, volume PGCT-3, pages 38–44; March, 1956.

described in the block diagram of Figure 13A and B.

Units of particular interest will be detailed in the following paragraphs.



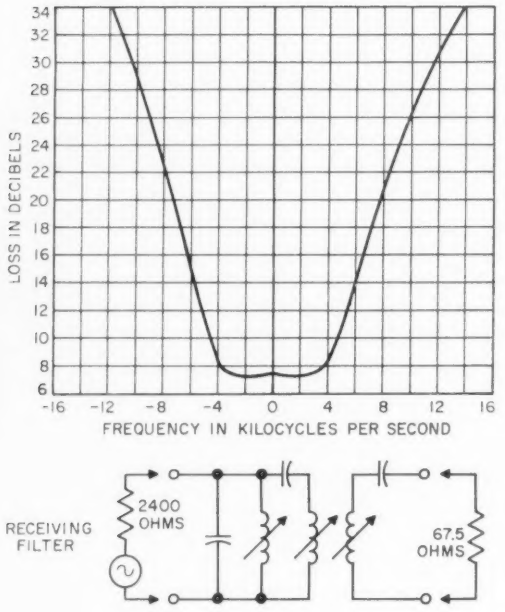


Figure 12—*K24A* channel-filter insertion-loss curves.

The carrier oscillator used in the transmitting section was described earlier. The remainder of the circuit consists of two passive elements, a balanced lattice modulator, and a transmitting filter of the type shown in Figure 11.

The receiver essentially consists of four elements: a variolosser controlled by the level of the received frequency-shift signal in the signaling unit, a carrier amplifier, a local oscillator, and a ring demodulator.

The ability of the receiver to detect double-sideband suppressed-carrier amplitude-modulation signals is due to a synchronizing circuit that controls the local oscillator. With the synchronizing circuit disabled the unit resembles a conventional nonsynchronous single-sideband demodulator. In the case of the reception of double-sideband suppressed-carrier signals, the synchronizing circuit detects the zero crossings of the double-sideband channel signal. These occur at twice the suppressed-carrier rate, with perturbation at a rate equal to twice the modulating frequency. The local oscillator is designed to synchronize at alternate impulses provided by the synchronizer and to ignore the perturbing impulses due to modulation. The result is a local oscillator source, synchronized rigidly in both frequency and phase, that therefore makes possible coherent detection of the double-sideband suppressed-carrier signal.

The receiving section of a channel modem is thus a fixed-frequency receiver that, with an appropriate transmitter, can in principle, detect double-sideband transmitted-carrier amplitude modulation, double-sideband suppressed-carrier amplitude modulation, or single-sideband. If any carrier is present in the received double-sideband signal, it is detected as direct current and can be used for control purposes.

The excellent phase linearity of the modem can be judged from Figure 14, which displays the response of the modem to square-wave excitation at various frequencies.

As noted earlier, the transmitter and receiver, on a common modem card, operate in two separate bands of frequencies. A number of considerations led to this arrangement. High-frequency operation of transmitters permits an economical crystal-oscillator design. On the

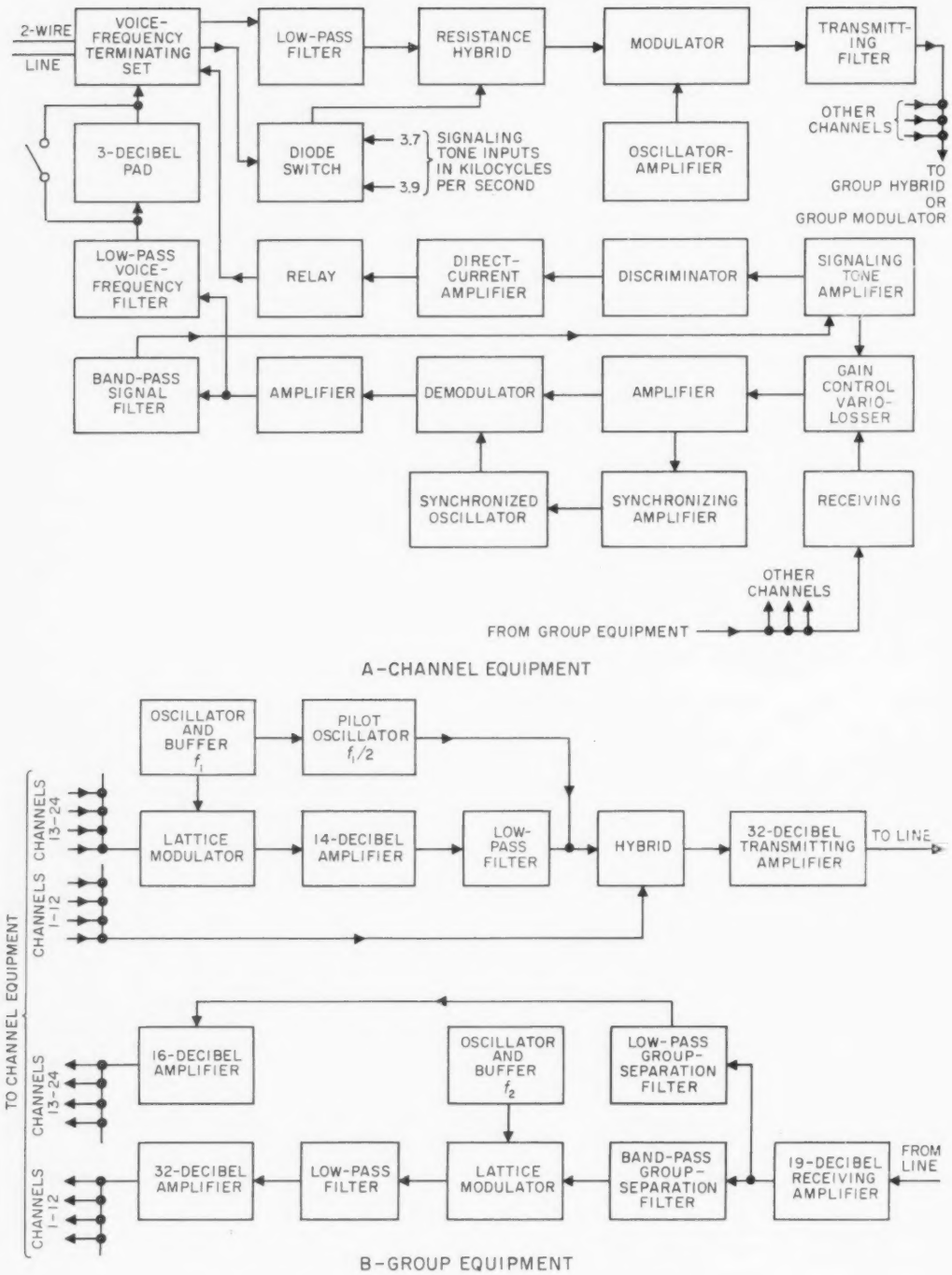


Figure 13—Channel and group equipment arrangement of *K24A* system.

other hand, a receiver operates better at low frequencies due to improved performance of transistors. In addition, of course, the wide

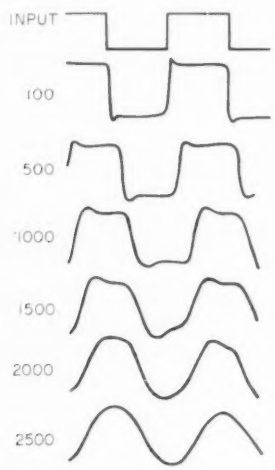


Figure 14—Channel response to square-wave input. The numbers indicate the frequency in cycles per second of the waveform at the demodulator output.

frequency separation obviates any coupling problems between transmitter and receiver on a single card.

2.2.2 Signaling Unit

There are a number of methods of transmissing switching information between exchanges. The simplest, and perhaps most common, is *E* and *M*, which is essentially the provision of a 4-wire low-speed data-transmission channel in connection with the voice circuit. Two other systems are usually referred to as loop and revertive. In loop signaling, the 4-wire *E*-and-*M* data circuit is converted to 2-wire operation over the loop used for voice transmission. An optional version provides supervisory information on a third control wire, eliminating the trunk circuit normally required. In revertive signaling applications, the functions of loop signaling are augmented by a third supervisory function. Space does not permit a detailed description of these systems.

To satisfy the functional requirements of these three systems of signaling transmission, three options of signaling unit are available as shown in Table 2.

The need for transmission of more than one bit of information in the case of revertive signaling influenced the choice of frequency-shift modulation for the signaling channel, in that three states are possible: one or the other of the two frequencies, or, for short durations, neither or both.

Transmission of the frequency-shift signal is accomplished by the provision of two tones for all 24 channels from a common supply. The tones are at 3.7 and 3.9 kilocycles per second, just above the voice-frequency band. The shift is achieved in the signaling transmitter by a diode switch designed to fade-in and fade-out the alternate frequencies and thus hold fortuitous switching distortion to acceptable limits. The signaling receiver is a conventional Round–Travis discriminator serving the dual purpose of driving a signaling relay and an automatic-gain-control rectifier.

Three steps of signaling distortion adjustment are provided to compensate varying external plant conditions.

2.2.3 Voice-Frequency Terminating Unit

This unit contains as its main element a transformer bridge, or hybrid, that converts the 4-wire voice circuit provided by the modem to a 2-wire circuit for connection to the exchange loop. It similarly permits the conversion of the 4-wire facility of the signaling unit into loop signaling.

The Sonad has been designed so that it can be added to the basic terminating card at any time. An optional version of the terminating unit contains a compandor, thus permitting the application of the system to exceptionally noisy transmission media.

Figure 15 indicates the channel response characteristics. It is largely influenced at the lower end of the spectrum by the characteristics of the terminating unit and, at the upper end, by the signaling-tone rejection filters in the signaling unit.

2.2.4 Repeaters

The main element of the repeater is the 32-decibel amplifier described earlier. The regulating repeater, shown in block-diagram form in Figure

16, contains an additional circuit. This circuit consists of a frequency-selective amplifier that picks off a pilot frequency located midway between the two 12-channel groups, a rectifier, and an adjustable-gain stage. Gain adjustment

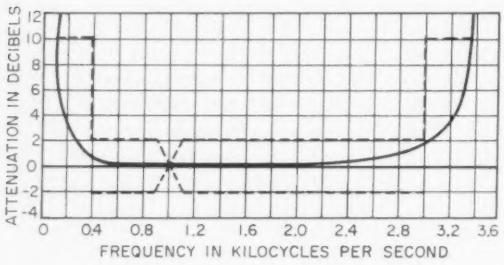


Figure 15—Channel characteristics for frequency and attenuation in decibels relative to response at 1 kilocycle per second. The dashed line is the toll objective.

is achieved by thermistor action in the collector-to-base feedback loop of a single-stage amplifier. The stage is a reflex circuit in that it serves both as a wide-band multichannel signal amplifier and as a direct-current amplifier for the control of its thermistor feedback circuit.

The repeater also accommodates, as plug-in elements, a span pad and a span equalizer. Power is obtained from a constant-current 56-milliampere source, series supplied from a phantom circuit formed by the east and west pairs of a system.

3. Applications and Maintenance

At the outset of the development of the *K24.1* system, it was determined that the success of 5-to-15-mile (8-to-24-kilometer) short-haul carrier would be very dependent on the simplification of application, installation, lineup, and maintenance techniques to a degree not previously required in the common-carrier field. How the system meets these objectives is discussed here.

The high density of circuits used in the exchange area of telephone plant, together with the obvious impracticability of having adequate staffs of skilled personnel available for its installation, adjustment, and maintenance, dictated the necessity of eliminating measurements and calculations normally associated with a carrier installation. The end product is a system that incorporates one continuously adjustable control per channel for the purpose of accurately setting up the circuit equivalent. In the remainder of the system, adjustments are provided on a group basis in discrete steps. These adjustments relate to span pads and equalizers at the input to both the repeaters and the receiving amplifiers of the terminals.

3.1 System Description

The *K24.1* system provides up to 24 circuits on two cable pairs between offices where the attenuation of these trunks does not exceed approximately 200 decibels at 480 kilocycles per second. This is a distance varying between 10 and 20 miles (16 and 32 kilometers), depending on the gage and capacitance of the cable pairs involved. The modulation plan of the system is shown in Figure 2.

The line frequencies for the channels of a terminal are on 16-kilocycle-per-second centers. Each channel occupies 8 kilocycles per second of spectrum, that is, 4-kilocycle-per-second upper and lower sidebands. The line frequencies for the channels at the opposite terminal are also on 16-kilocycle-per-second centers, but all frequencies are shifted 8 kilocycles per second above those for the first terminal. Since each channel occupies 8 kilocycles per second of spectrum, there is no overlapping of frequencies when considering the two directions of transmission. This staggering of frequencies for opposite directions of transmission eliminates near-end-crosstalk problems and simplifies filter requirements.

The lowest channel carrier frequency in the top 12-channel group is 292 kilocycles per second, and the highest channel carrier frequency is 476 kilocycles per second.

The line frequencies of the high group were chosen to permit operation of 12-channel systems in the same cable sheath with Western Electric Company *N-1* carrier systems.

The system terminal is designed to mount in 31.5 inches (80 centimeters) of vertical space on 19-inch (48-centimeter) relay racks. Four system terminals will mount in one 11-foot 6-inch (350-centimeter) bay.

The system uses transistors only, with unit

plug-in module-type construction employed in both terminal and repeater equipment. Printed-circuit wiring techniques are used on all plug-in cards. Printed wiring connectors on the ends of each card mate with multicontact receptacles mounted on the terminal shelves. The boards are guided into position by grooves in the shelves.

Handles on each plug-in unit assure easy withdrawal of the board from the shelf. There are 7 shelves in a terminal, the top shelf housing the common equipment and the remaining 6 shelves, the 24 channel terminals. The circuit for each channel is divided into three cards; modem, signaling, and voice-frequency termination. Three shelves mount 12 channels, with the three cards of any one channel appearing vertically in the terminal.

A separate card is used for each of the three available signaling options; loop, *E* and *M*, and revertive.

Compandors and Sonads are available on an optional basis and, when furnished, are included on the voice-frequency termination cards. A Sonad can be added to a channel at any time by plugging the subassembly into the termination card of the channel. Circuit interconnections are made by a multipin plug-and-socket arrangement. Compandors can also be added to a channel at any time, merely by changing the termination card; no wiring or strapping procedures are required to implement the change.

A static 48-to-130-volt direct-current/direct-current converter completes the equipment available at terminal locations. Appliqué auxiliary line equipment such as filters, terminations, matching transformers, et cetera, normally associated with carrier applications, are not required when using this equipment. For instance, when the loop signaling option is used, step-by-step equipment, or the outgoing trunk circuit (depending on whether the 3-wire or the 2-wire loop signaling unit is to be used) connects directly into the drops of one terminal, and the opposite terminal drops are connected directly into the incoming selector equipment.

Regulating and nonregulating repeaters are used with this system. Both types of repeaters furnish a normal 25 decibels of gain at the highest line frequency of 480 kilocycles per second. The regulating repeaters compensate for line variations of ±8 decibels. Every third repeater in a system should be of the regulating type. Repeaters are installed at the same point for both directions of transmission, regardless of the type required. Repeaters are power-supplied over the same pairs on which they are installed and up to five, in tandem, may be supplied from each end. A system may use up to 10 repeaters, therefore, without requiring an intermediate power source.

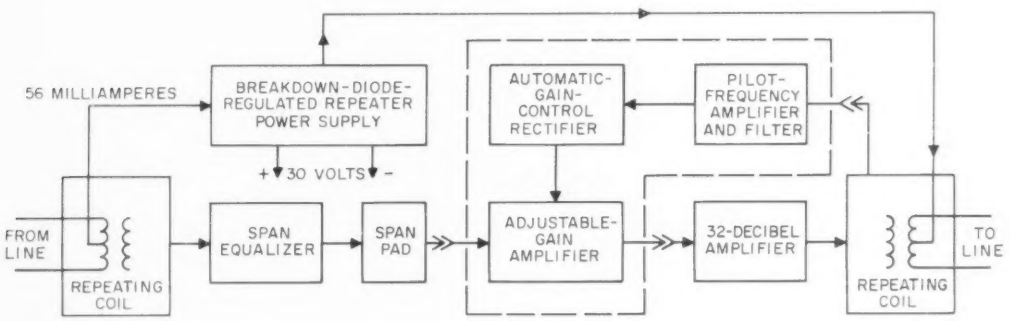


Figure 16—Regulating and nonregulating repeater. For the nonregulating repeater, the items within the dashed outline are omitted and the span-pad output is strapped directly to the 32-decibel-amplifier input.

3.2 Installation Engineering

A minimum of engineering is required to install this system. At terminals, it is only necessary to specify the number of channels and the type of signaling option desired. The number and type of repeaters are determined by referring to Table 3. It will be noted repeaters are normally

located at 6000-foot (1829-meter) (*H*-loading) intervals, or multiples thereof, depending upon the gage and capacitance of the cable pairs involved.

If voice-frequency-loaded pairs have to be unloaded for an initial installation of this carrier system, it is possible to effect a considerable economy in the cable-splicing operations involved. This is accomplished by unloading more than the number of pairs required for initial installation. These extra pairs are then reloaded by mounting the loading coils on plug-in cards designed to fit in the repeater housings. Additional systems are then added by merely replacing the loading-coil cards with repeaters.

TABLE 3

K24A Repeater Spacing

Cable		Repeater Spacing	
Gage	Manufacturers' Designation	Feet	Meters
16	*Toll*	12 000	3 658
16	*NH*	12 000	3 658
19	*Toll*	12 000	3 658
19	*DNB*	6 000	1 829
19	*CNB*	6 000	1 829
22	*Toll*	6 000	1 829
22	*TS*	6 000	1 829
22	*CSA*	6 000	1 829
24	*NH*, 0.065	6 000	1 829
24	*CSH*, 0.072	6 000	1 829
24	*DSH*, 0.084	4 500	1 372

3.3 Installation, Alignment, and Operation

3.3.1 Repeaters

A repeater alignment test set is the only instrument required when aligning or checking the performance of a repeater. Operating power for this set is derived from mercury batteries. When installing repeaters, the instrument is calibrated so as to indicate easily the appropriate span pad and equalizer to be used. No other adjustments or measurements are required for installing either regulating or nonregulating repeaters.

3.3.2 Terminals

Alignment of the terminals consists essentially of selecting the appropriate equalizer and span pad and adjusting the circuit equivalents. The alignment test set is used at the terminal to select the pads required in a fashion similar to that used at the repeaters.

Extension units provide access to units permitting measurements on both a bridging and a terminating basis.

3.3.3 Signaling

An infrequently required, but nevertheless available, step adjustment is the one for pulse distortion. By a solder-strap change provided, bias can be varied ± 10 percent from the nominal setting.

3.3.4 Alarm System

The design of the alarm system aimed at three objectives: (A) Protection of the equipment from damage due to abnormal power-supply occurrences. (B) Prevention of seizure of the exchange facility in the event of equipment failure. (C) Indications that would pinpoint as closely as possible the source of the trouble.

The alarm system is shown in block form in Figure 17.

A predetermined change in the voltage of either the exchange battery or the output of the power-supply regulator actuates a relay and shuts down the terminal.

The relay can also be actuated by the failure of the system pilot or the failure of loops in all of either the lower automatic-gain-control group of 12 channels, the upper group of 12 channels, or all 24 channels.

Measurements at the test jacks associated with the alarm system allow the trouble shooter to isolate the failure in one of the following areas:

(A) Power supply.

(B) Line transmission facility, including the line amplifier in the terminal, the repeaters, or the line itself. A further degree of identification of the offending unit requires the use of the alignment test set.

(C) Group modulators, demodulators, and signal supply.

Protection of the exchange from tie-up due to system failure is achieved by the inherent

operating characteristics of the system. The signaling conditions corresponding to channel failure are those of either battery-normal or no-battery at the originating office and loop-open at the terminating office. This assures that in the case of failure the exchange equipment and subscribers are released.

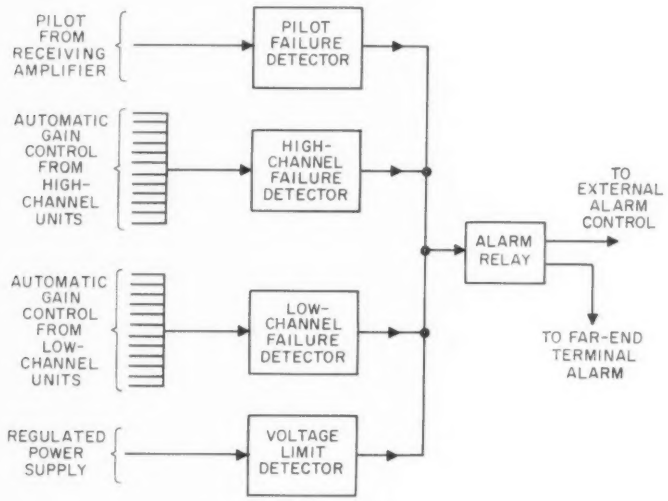


Figure 17—*K24.1* terminal alarm.

3.3.5 Trouble Shooting and Maintenance

Virtually complete trouble shooting can be accomplished by measurements made through the unit extension panels, or by the use of the alignment test set. Test points have been eliminated from this equipment in the interest of economy. However, the extension panels furnish even-more-complete access to each unit than that normally provided by test points.

The alignment test set described earlier is equipped, of course, to provide all of the pertinent information necessary to indicate the performance of any repeater.

3.3.6 Power Supply

The *K24.1* system operates from 30 volts direct current. The supply is derived from the 48-volt exchange battery at the terminal locations through a regulator. The average total consumption of a 24-channel terminal is approximately 100 watts at 48 volts.

The repeater requirement is 30 volts at 56 milliamperes. Repeaters are supplied in series, and the individual supply is regulated. Accordingly, a 5-repeater system requires a source supply of 150 volts plus the voltage drop of the cable pair. Thus, for a 1-repeater system, the 48-volt exchange battery is an adequate supply. For two repeaters or more, the 48-volt and 130-volt exchange supplies must be put in series. When 130-volt battery is not available in an exchange, the 48- to 130-volt direct-current converter may be used.

3.4 Other Applications

The favorable multichannel loading characteristics necessary for the efficient utilization of transistor repeaters in the exchange application make the system applicable to any transmission medium where signal power capacity is limited. The system can therefore be applied to radio links without adaptation of any kind.

Versions of the *K24.1* providing higher channel capacities and closer channel spacing for use on microwave links are under development. Others under development are for medium-haul cable applications.

High-Speed Transmission of Numerical Data Over Telephone Channels

By ANATOLE GIRINSKY and PAUL ROUSSEL

Compagnie Générale de Constructions Téléphoniques; Paris, France

ORGANIZATIONS such as those providing electricity, transportation, and military protection have operating units distributed over wide geographical areas that must report significant changes in conditions to a supervisory headquarters and respond quickly to its orders. For this purpose it may be found desirable to utilize existing communication channels that offer sufficient reliability. It may, however, be necessary to encode information, often in numerical form, to reduce transmission time and to permit its direct utilization by both the supervisory system and the operating units. Further economy in transmission time can be obtained by the temporary storage of information at each end of the circuit, particularly if a sequence of operations must be supervised.

In the equipment to be described, the information is transmitted in the form of binary digits or bits. The conversions from the original form to that suitable for transmission and vice versa are accomplished by terminal equipments that adjust the speed of transmission to the characteristics of the transmission network. The storage unit at the receiver records the data for direct utilization or converts it into some other form if that is necessary.

There is no direct-current component in the signaling code that would be lost in passing through a repeater or transformer. In addition, the receiver is automatically synchronized by the transmitter to provide good reliability and ready adaptability to all types of telephone systems.

1. Principles

The information elements are transmitted sequentially, one at a time, and are transmitted during equal time intervals. The transmission network will be maintained at all times in either of two conditions.

1.1 Code

The condition of the line will be in accordance with the following three rules.

(A) The condition of the line is reversed at the beginning of each time interval.

(B) If the binary digit to be transmitted in that time interval is *1*, the line condition will be changed halfway through the time interval from that to which it was changed at the beginning of the time interval.

(C) If the binary digit is *0*, the line is maintained for the duration of the interval in the condition to which it was changed at the beginning of the time interval.

It is evident that the condition of the line has no real meaning as the same bit might be transmitted in either condition, depending on the condition of the line at the start of transmission.

FORMS	*0*	*1*		*0*	*1*		*1*		*0*	*0*
FIRST	*A*	*B*	*A*	*B*	*A*	*B*	*A*	*B*	*A*	*B*
SECOND	*B*	*A*	*B*	*A*	*B*	*A*	*B*	*A*	*B*	*A*

Figure 1—Any group of binary digits may be indicated by two sets of signals depending on whether the line is in condition *A* or *B* at the start of transmission. The condition is changed at the start of each interval and remains constant for *0* but changes after half an interval for *1*.

Examples of two ways of transmitting the same coded information, *0101100*, are shown in Figure 1, where the two conditions are designated as *A* and *B*.

1.2 Modulation

There are two basic types of transmission, which may be designated as unmodulated and modulated. The former would correspond to two potentials of opposite polarities.

Modulation may take one of several forms. In all of them, one condition will correspond to the absence of any signal and the other condition will be designated by a sine wave containing two or more cycles in each transmission interval. In the first case, the highest signaling speed over the line may be obtained if the carrier frequency is suitably chosen. These signals are represented

in Figure 2. There is no direct-current component in the signaling code that would be lost in passing through a repeater or transformer. In addition, the receiver is automatically synchronized by the transmitter to provide good reliability and versatility.

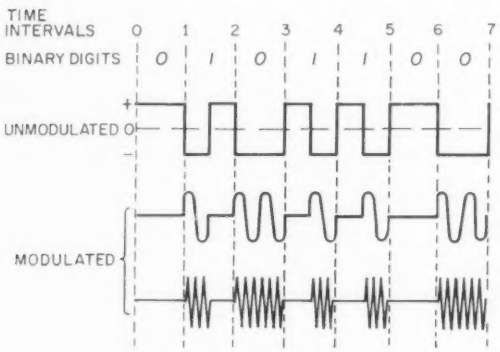


Figure 2—Three ways of transmitting binary digits. Each digit is transmitted in the time interval in which it is shown and in accord with Figure 1. In the unmodulated case, the polarity of a direct voltage is reversed to indicate the changes in condition. In the two examples of modulation, only the number of sine waves in the time interval differs.

2. *Transmission*

The conversion of a sequence of binary digits into coded signals requires 3 bistable trigger circuits, *A*, *B*, and *C*. The speed of operation is controlled by a pulse generator producing two series of pulses, *P1* and *P2*, having the same time intervals between pulses but being offset from each other by half this time interval.

Figure 3 shows the development of a coded signal for the binary series *0101100*. The generated pulse train *P1* determines the time intervals for the output signal. *P2* controls the half intervals at which the condition of the line is changed for a *1*.

Trigger circuit *A* is controlled by *P2* and by the input digit. If the input digit is *1*, it produces output for the full transmission interval set by *P2*. No output occurs for a digit *0*.

Trigger circuit *B* is controlled by *P1* and trigger *A* and will change its condition only if *A* has no output when a *P1* pulse occurs.

Trigger *C* is connected to *P1* and *P2* and in the absence of trigger *B* would change its condition at each half interval when these pulses are received. It is also under control of trigger *B* and will maintain its condition for a full interval of *P1* immediately after a change in output from trigger *B*. The output of trigger circuit *C* is the coded signal for transmission to the line.

3. *Reception*

The received signal will be distorted and weakened so it is amplified and reshaped by trigger *D* of Figure 4. Trigger *D* controls a pulse generator *P3* that produces a short pulse each time there is a transition in its output. Each of these pulses resets another pulse generator *P4* that produces a pulse only if it is undisturbed for $\frac{3}{4}$ of the time interval assigned for the transmission of each digit. These pulses, therefore, correspond to receipt of the binary digit *0*.

Trigger *E* changes its condition for each pulse received from *P3*. In addition, it is reset to *0* for each pulse from *P4*. It starts each time interval in condition *1*. If the received digit is *1*, *E* is reset to *0* at the half interval under control of *P3*. If the digit is *0*, *E* is reset at the beginning of the last quarter interval under control of *P4*.

The synchronizing pulses *P5* that control the receiver are obtained from trigger *E*. They are derived from the rising pulses that occur when

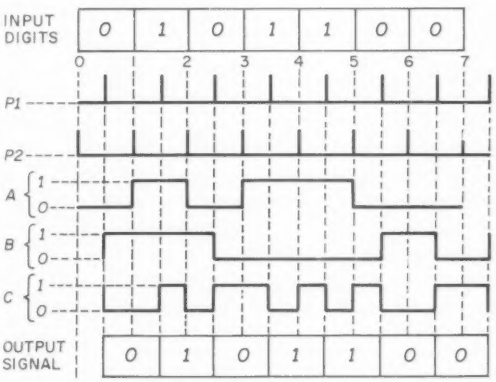


Figure 3—Development of coded signals. *P1* and *P2* are two series of timing pulses. Trigger *A* produces output during interval of *P2* if digit *1* is to be sent and no output for *0*. Trigger *B* changes its condition if *A* provides no output when *P1* pulses occur. Trigger *C* reverses its output under control of interleaved pulses from *P1* and *P2* and is held without reversal for a full interval of *P1* if a change has occurred in *B*.

E is reset to condition *1* at the start of each time interval and are delayed to occur at the half-time intervals of the period assigned to the transmission of each digit.

Trigger F changes its condition under control of *P4*, which is at the beginning of each last quarter of the time interval when a *0* is being received.

Under control of *P5*, trigger G assumes the same condition as trigger F. Thus at the half-time intervals corresponding to *P5*, G is in the same condition that F had at the start of the time interval. Also, for each pulse *P5*, H goes to *0* if triggers F and G were in different conditions a half-interval before or to *1* if they were in the same condition. H therefore produces a decoded version of the received signal delayed by 1.5 time intervals.

1. Experimental Equipment

4.1 Transmitter

The experimental transmitting equipment, outlined in Figure 5, makes use of a start-headed 5-element code. Timing is under control of a generator that produces a sine wave of suitable frequency and by means of other circuits develops from it a square wave of the same frequency and pulse trains *P1* and *P2*.

There are 16 inputs, *0–15*, to the pyramidal diode matrix with inputs *3* through *7* connected to switches *M1–M5* to permit manual setting of each at either *0* or *1*. Input *2* is the start element and input *8* may be set manually to provide an additional start element enabling the transmission of two consecutive codes. The next transmitted elements will all be *0*'s since those subsequent to terminal elements are open in this experimental version.

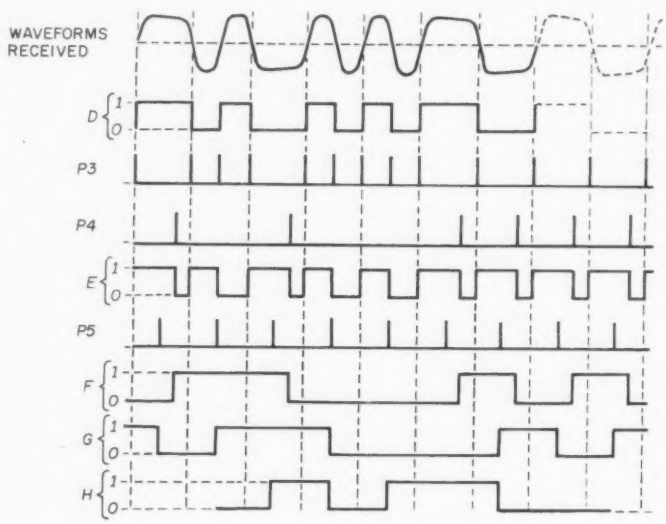


Figure 4—The received signal is reshaped by trigger D. Pulses *P3* occur at each transition of D. Pulses *P3* reset pulse generator *P4*, which then produces a pulse only if it is not reset for $\frac{3}{4}$ of a time interval, its pulses corresponding to binary digit *0*. *P4* pulses return E to zero and *P3* pulses change its condition. Synchronizing pulses *P5* correspond to rising pulses from E with a half-interval delay. Trigger F changes its condition at each pulse from *P4*. *P5* causes G to change halfway during the interval to the condition of F at the start of the interval. H at *P5* timing goes to *0* if F and G were different a half-interval before and to *1* if they were the same. H is the decoded received signal delayed 1.5 time intervals.

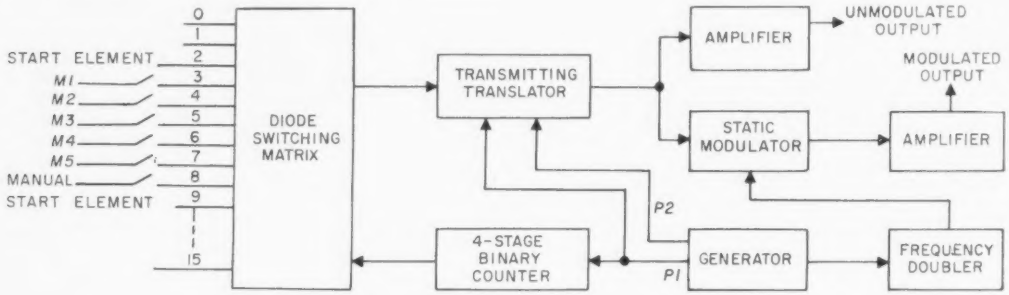


Figure 5—Transmitting system.

The 16 inputs are connected sequentially to the transmitter translator by the 4-stage binary counter (count of 16), the pulses from which control the polarity of bias on the diodes in the pyramidal matrix to provide this switching. Switching is continuous under synchronizing control from the generator.

The transmitting translator provides the functions discussed in section 2. It produces unmodulated signals that are amplified and available for transmission. This output also goes to a static modulator to key a sine wave of twice the frequency of the generator. After amplification, this signal is available for transmission.

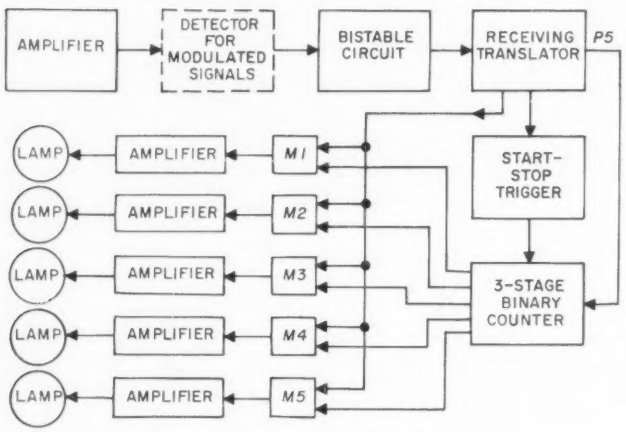


Figure 6—Receiving system.

The time interval for the transmission of each binary digit depends on the frequency of the generator and its corresponding frequency doubler. Plug-in units for these two functions provide for variation of frequency. All remaining circuits are aperiodic and need not be modified to operate at other time intervals. Intermediate transmission speeds may be obtained by using an auxiliary adjustable-frequency oscillator. Table 1 shows various transmission speeds for the line input equipment.

TABLE 1

TRANSMISSION SPEEDS

Transmission Speed					
Bauds	4000	3000	2000	1000	500
Bits Per Second	2000	1500	1000	500	250
Time Interval Per Binary Digit in Microseconds for Unmodulated Signals	500	666	1000	2000	4000
Frequency in Cycles Per Second for Modulated Signals	—	—	2000	1000	500

4.2 Receiver

The signals received from the line go to the amplifier shown in Figure 6 and are also reshaped. If modulated signals are being received, the detector circuit is used. The circuit following the amplifier is bistable and is triggered for each change in polarity of the signal wave. The receiving translator performs the functions outlined in section 3.

The receiving translator derives from the incoming signal not only the binary digits corresponding to the information to be transmitted but also a series of pulses, *P5*, that occur at the beginning of each time interval assigned to a digit.

If no signals are being received or if the signals correspond to *0*, the start–stop trigger is in the stop condition and the binary counter is blocked so that no output occurs to the group of storage circuits *M1–M5*. On receipt of a *1*, which is the start signal, the start–stop trigger operates and permits the *P5* pulses to actuate the 3-stage binary counter. These pulses are distributed sequentially over the *M1–M5* circuits, each thereby being enabled in turn to accept its binary digit from the receiving translator. The start–stop trigger is then reset to the stop condition.

A fault counter is provided in the receiver and records each time that the indicated character differs from the one manually marked.

The receiver operates at the same speed as the transmitter, its synchronizing pulses being derived from the transmitted signal. Those circuit elements that must be changed when shifting between modulated and unmodulated signals or from one transmission speed to another are built as plug-in units.

The sensitivity of the receiver is such that it will operate on signals that are 4.5 nepers (39 decibels) below the normal output power of the

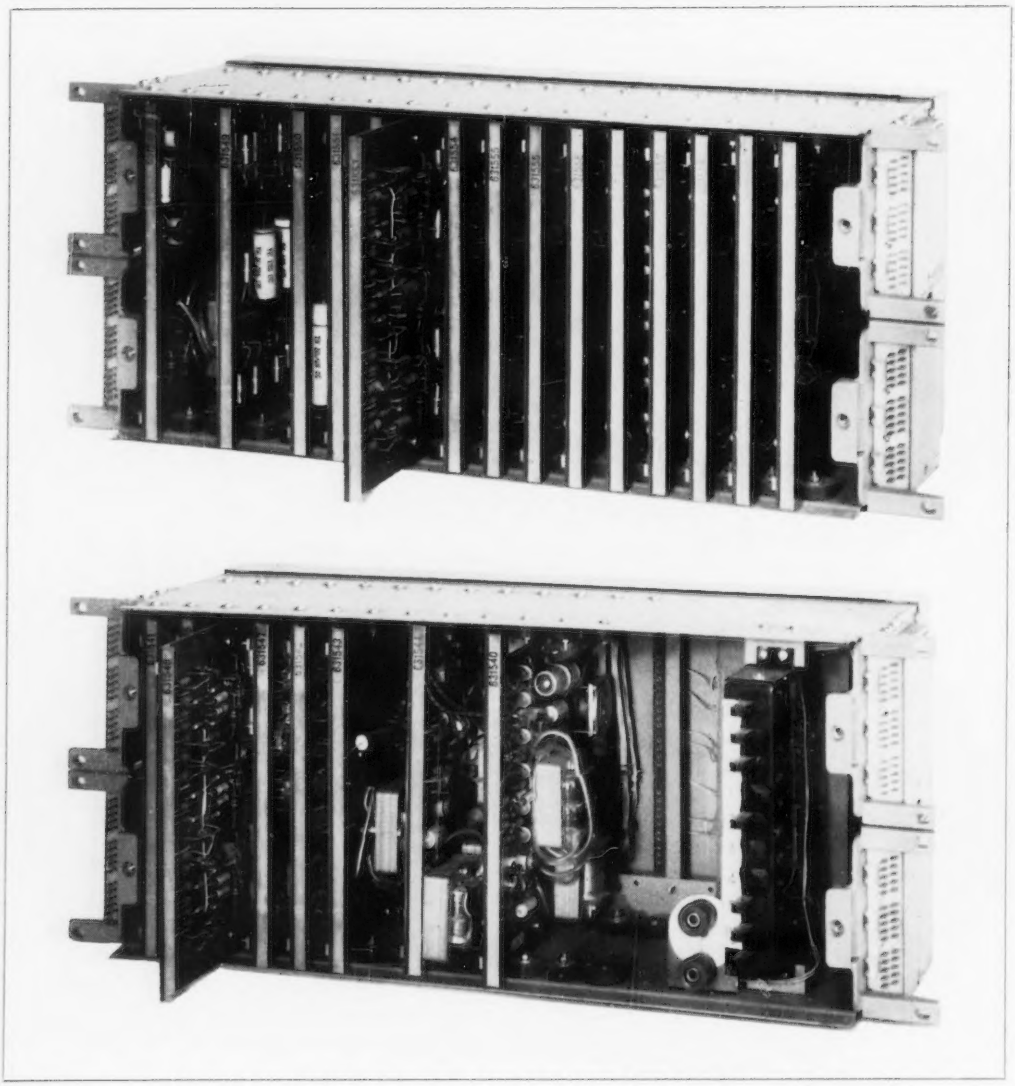


Figure 7—At top is the transmitter assembly and the receiver is below it. The printed-wiring panels plug in.

transmitter. If the signal drops to 5.5 nepers (48 decibels) below the transmitter level, it will produce no response.

4.3 Construction

As will be seen in Figure 7, the components are mounted on printed-circuit panels that are plugged into an assembly frame. Transistors are used throughout.

5. Tests on Loaded Cable

Tests were made on two lengths of loaded cable of type *H88*. One length was of 100 kilometers (62 miles) with one repeater and the other was of 264 kilometers (164 miles) with 6 repeaters. Each circuit was a loop and employed 2 pairs in the same cable. Repeaters were of the 4-wire type and placed at 45-kilometer (28-mile)

intervals. The *H88* cable differs from the usual types in that transformers are inserted at approximately 9-kilometer (5.6-mile) intervals. Loading coils of 88-millihenry inductance are inserted every 1830 meters (1.14 miles).

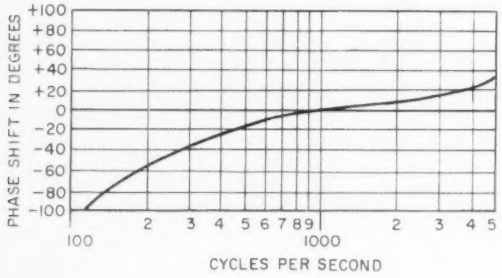


Figure 8—Phase characteristic of repeaters.

5.1 Cable Characteristics

The individual wires have a diameter of 0.9 millimeter (approximately 19 American Wire Gage) and the following properties were measured.

Resistance = 54.5 ohms per kilometer
= 87.7 ohms per mile

Capacitance = 38.5 nanofarads* per kilometer
= 62.0 nanofarads per mile

Inductance = 0.65 millihenry per kilometer
= 1.05 millihenries per mile.

5.2 Transformer Characteristics

Measurements made on the transformers show they may be represented approximately by an inductance of 10 henries in parallel with a resistance of 50 000 ohms.

5.3 Repeater Characteristics

The repeaters are of the 4-wire *PTT* type. Compensation for distortion is provided by reverse feedback and the values and arrangement of the capacitors and resistors in the feedback circuit can be modified to permit the over-all repeater characteristics to be changed.

The phase characteristics of a repeater are shown in Figure 8. The measurements were made with the repeater adjusted to give the same gain over the amplified section between 200 and 2400 cycles per second.

5.4 Over-All Channel Characteristics

The main characteristics of the over-all transmission system including loading and repeaters on the 264-kilometer (164-mile) cable, are given in Figures 9, 10, and 11. The impedance and attenuation curves, Figures 9 and 10, are comparable to those usually encountered with such cables but the phase characteristic of Figure 11 shows important distortion at both low and high frequencies.

The distortion at the high-frequency region is due to the proximity of these frequencies to the cutoff frequency of the loaded cable.

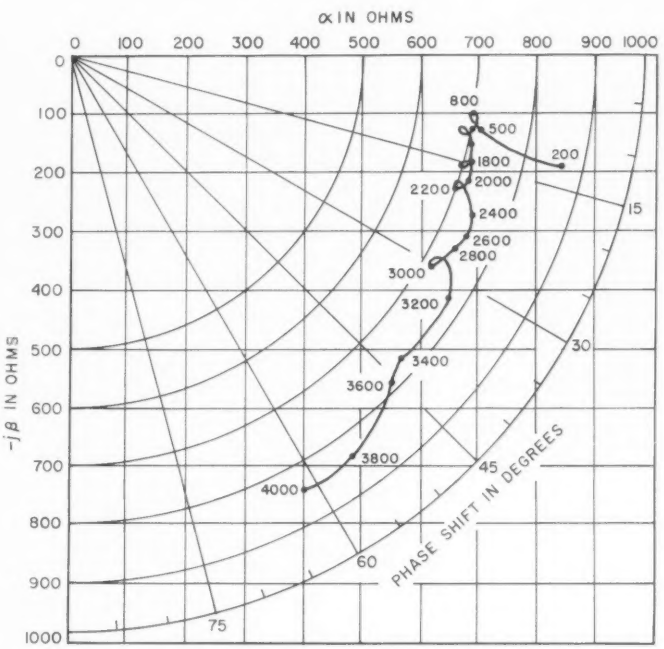


Figure 9—Impedance values for the loaded 264-kilometer (164-mile) cable. Frequency indications are in cycles per second. $Z = \alpha + j\beta$ versus frequency.

* Nanofarad = 10^{-9} farad.

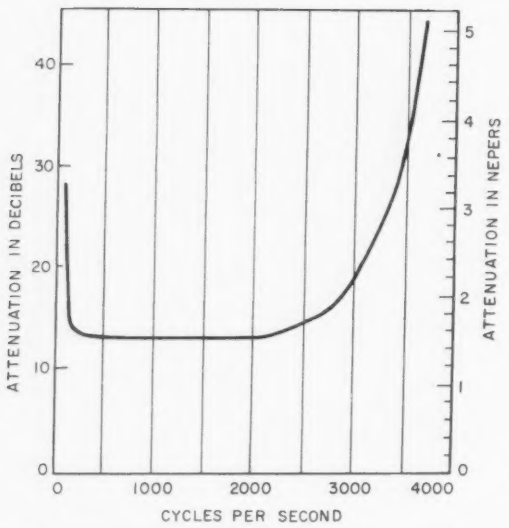


Figure 10—At the left is the attenuation as a function of frequency for the 264-kilometer (164-mile) cable.

The low-frequency distortion is not normally of importance. The phase shift as a function of frequency of the normal *H88* line may be obtained from the equation[1]

$$\beta = \omega(LC)^{1/2} f(x)$$

where

$$f(x) = (1/2^{1/2})\{1 - x^{-2} + [1 + 2(m^2 + n^2)x^{-2} + x^{-4}]^{1/2}\}^{1/2}$$

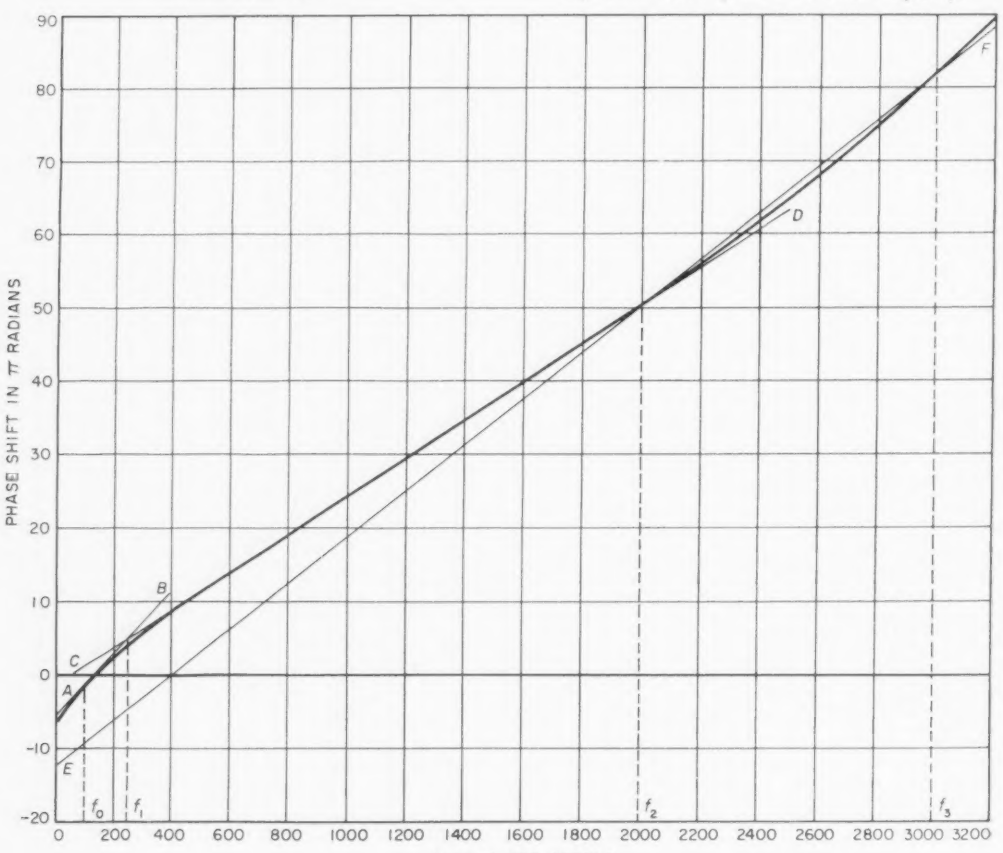
with R, L, C, and G being the line constants and

$$m = \frac{1}{2}\left[\left(\frac{RC}{LG}\right)^{1/2} + \left(\frac{LG}{RC}\right)^{1/2}\right];$$

$$n = \frac{1}{2}\left[\left(\frac{RC}{LG}\right)^{1/2} - \left(\frac{LG}{RC}\right)^{1/2}\right];$$

[1] E. A. Guillemin, "Communication Networks," John Wiley, New York, 1931–1935, volume 2; page 84.

Figure 11—Below, phase shift versus frequency.



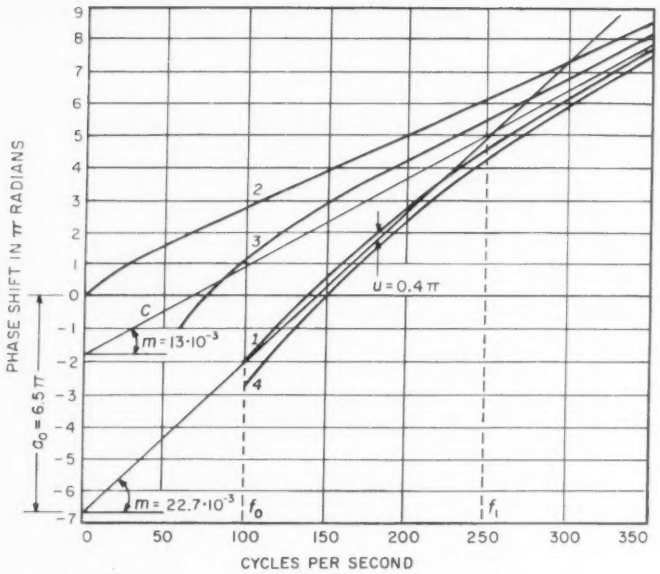


Figure 12—Phase shift at the lower frequencies. Curve *1* was measured for the 264-kilometer (164-mile) circuit under test. Calculated curves are given for (*2*) normal loaded *H88* cable, (*3*) *H88* cable with transformers, and (*4*) cable, transformers, and repeaters.

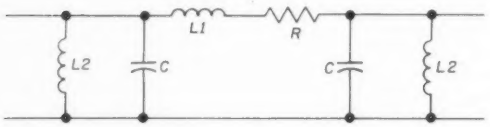


Figure 13—Equivalent circuit at low frequencies for a 9.15-kilometer (5.7-mile) section of *H88* cable with loading. $C = 0.175$ microfarad, $L1 = 0.44$ henry, $L2 = 20$ henries, and $R = 500$ ohms.

and

$$x = \frac{\omega}{\left(\frac{RG}{LC}\right)^{1/2}}.$$

Curve *1* of Figure 12 is the measured phase characteristic of the line under test and curve *2* is calculated as above for a normal *H88* line of 264-kilometer (164-mile) length. The difference is due in part to the effect of the transformers cutting the line into 9-kilometer (5.6-mile) sections.

Assuming for convenience that these sections are 9.15 kilometers (5.7 miles) long, which is equivalent to 5 loading-coil spacings, at low frequencies each such section may be regarded as being a network of lumped constants since its length is small compared to the wavelength (about 66 kilometers at 300 cycles per second). Each length may be approximated by the network given in Figure 13.

The transfer factor θ of such a network is given by

$$\exp \theta = \exp (\alpha + j\beta)$$

$$= 1 + \frac{p}{q} + \left[\frac{p}{q}\left(\frac{p}{q} + 2\right)\right]^{1/2}$$

where

p = impedance of the series arm

q = impedance of the parallel leg.

The imaginary part β of this transfer function represents the phase shift, which may be calculated for each frequency. Curve *3* of Figure 12 is such a plot.

If the phase shift contributed by the 6 repeaters is added to that of the line, curve *4* of Figure 12 results and corresponds well to the measured values shown in curve *1*.

6. Experimental Results

Figure 14 shows the circuit arrangement used in the experimental testing of the system. Operating margins were determined for minimum signal providing acceptable reception and for the maximum signal that could be tolerated by the receiver.

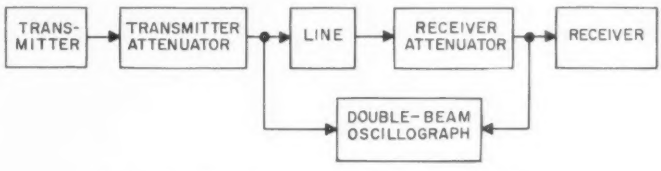


Figure 14—Circuit arrangement for experimental tests.

To determine the minimum signal, the transmitter attenuator was adjusted to give a prescribed output to the line and the receiver attenuator was adjusted to provide the weakest signal that gave satisfactory operation.

For the maximum permissible signal, the receiver attenuator was first reduced to zero attenuation and then, if necessary, the transmitter attenuator was adjusted until the upper limit of the receiver input level was reached.

6.1 Test on 100-Kilometer (62-Mile) Line

The 100-kilometer (62-mile) length of *H88* cable was provided with one repeater at the middle of the line and this was adjusted to reduce the attenuation to 1.6 nepers (13.9 decibels) between 200 and 2400 cycles per second. Table 2 shows the results.

6.2 Test on 264-Kilometer (164-Mile) Circuit

As previously stated, there were six repeaters in the 264-kilometer (164-mile) cable and its attenuation characteristic is given in Figure 10. The waveform of the received signal was improved by placing a reactance at the receiver input, the value of the reactance being determined experimentally for each type of transmission. The margin figures are given in Table 3.

TABLE 2

Tests on 100-Kilometer (62-Mile) Channel

Transmission Type	Carrier Frequency in Cycles Per Second	Transmission Speed		Operating Margin	
		Bauds	Bits per Second	Nepers	Decibels
Unmodulated	—	3000	1500	2.95	25.6
Modulated	500	500	250	3.75	32.8
Modulated	1000	1000	500	3.35	29.1

TABLE 3

Tests on 264-Kilometer (164-Mile) Channel

Transmission Type	Carrier Frequency in Cycles Per Second	Transmission Speed		Operating Margin	
		Bauds	Bits per Second	Nepers	Decibels
Unmodulated	—	2000	1000	2.65	23.0
Modulated	500	1000	500	3.40	29.6
Modulated	1400	700	350	3.00	26.1

Figure 15 shows reproductions of three oscillograms of different signals transmitted over the 264-kilometer (164-mile) channel in *H88* cable.

Phase distortion is noticeable at both the low and high ends of the transmission frequency band as indicated in Figures 12 and 16. It is an important factor in setting the operating limits of the system.

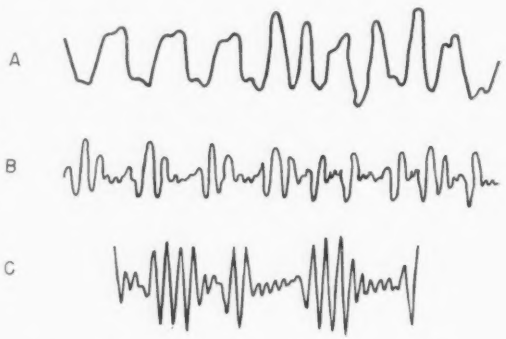


Figure 15—Reproductions of oscillograms of signals sent over the 264-kilometer (164-mile) channel of *H88* cable. *A* is an unmodulated signal at 2000 bauds for code *00000001101010O*. *B* is a modulated signal with a 1000-cycle-per-second carrier at 1000 bauds for the same code as *A*. *C* is a modulated signal with a carrier frequency of 1400 cycles per second at 700 bauds for code *101000*.

7.1 Effect on Pulse Form

7.1.1 General Features of Transmission Channel

The phase characteristic $\varphi(\omega)$ shown in Figure 11 differs much from the ideal, which would be represented by a straight line from the origin. Consequently, the received pulses are distorted, the degree of distortion being dependent on the transmission characteristic shown in Figure 10. The characteristic is comparable to that of an ideal band-pass filter in which

$$A(\omega) = 1 \quad \text{for} \quad \omega_0 < \omega < \omega_2$$
$$= 0 \quad \text{for} \quad \omega < \omega_0 \quad \text{or} \quad \omega > \omega_2.$$

Although the effects of differences between the ideal filter and networks obtained in practice have been examined[2] this difference is not important in this particular case and may be neglected in the following discussion.

[2] A. E. Sunde, "Theoretical Fundamentals of Pulse Transmission," *Bell System Technical Journal*, volume 33, pages 721–788; May, 1954; and pages 987–1010; July, 1954.

7.1.2 General Problem

Let us assume a signal $f(t)$ represented as a Fourier integral.

$$f(t) = \int_{-\infty}^{+\infty} g(\omega)\,[\exp\,(j\omega t)]d\omega.$$

If this signal is applied to a transmission line having an amplitude characteristic $A(\omega)$ and a phase characteristic $-\varphi(\omega)$, the response at the receiving end of the line is

$$S(t) = \int_{-\infty}^{+\infty} A(\omega)\,g(\omega)\,\{\exp\,\{j[\omega t - \varphi(\omega)]\}\}d\omega.$$

If $f(t)$ has a unit-step function

$$f(t) = \frac{1}{2} + \frac{1}{\pi}\int_0^{\infty} \frac{\sin \omega t}{\omega}\,d\omega$$

the response signal becomes

$$S(t) = \frac{A(0)}{2} + \frac{1}{\pi}\int_0^{\infty} \frac{A(\omega)\,\sin\,[\omega t - \varphi(\omega)]}{\omega}\,d\omega.$$

Generally, direct current is not transmitted through the circuit so that $A(0)/2 = 0$.

7.1.3 Linear Phase Not From Origin

If the phase shift is linear with frequency but is not zero at the origin, it is given by

$$\varphi(\omega) = m\omega - a_0$$

where

a_0 = phase shift for zero frequency

m = slope of curve in radians per angular velocity unit.

Assuming $A(\omega) = 1$ for all frequencies, the received signal becomes

$$S(t) = \frac{1}{2} + \frac{1}{\pi}\int_0^{\infty} \frac{\sin\,(\omega t - m\omega + a_0)}{\omega}\,d\omega$$

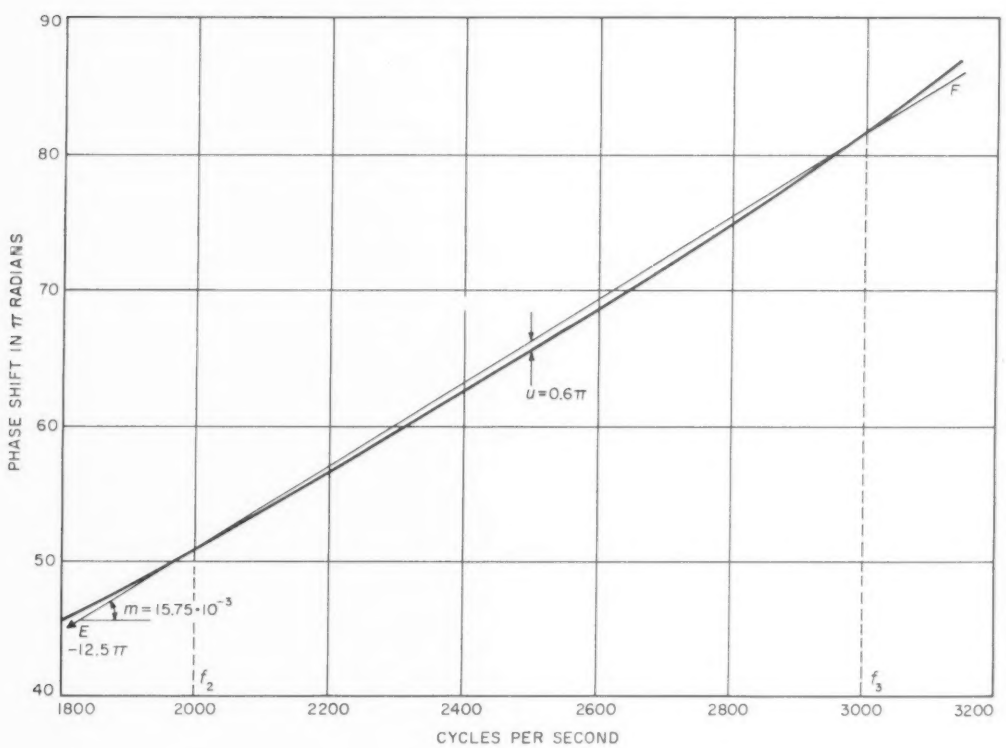


Figure 16—Phase shift at the higher frequencies for the 264-kilometer (164-mile) channel in *H88* cable.

or

$$S(t) = \frac{1}{2} + \frac{1}{\pi}\int_0^{\infty} \frac{\sin \omega(t-m)\cos a_0 + \cos \omega(t-m)\sin a_0}{\omega}\, d\omega$$

that is

$$S(t) = \frac{1}{2} + \frac{1}{\pi}\{\cos a_0 [Si\, \omega(t-m)]_0^{\infty} + \sin a_0 [Ci\, \omega(t-m)]_0^{\infty}\}.$$

The factor $\cos a_0$ represents the response signal from a circuit having the same phase characteristic but with the slope starting from the origin.

The factor $\sin a_0$ represents a spurious signal having maximum effect for

$$a_0 = (2K+1)\pi/2, \quad K = \text{any integer.}$$

For the network of section 7.1.1 having an amplitude characteristic limited to frequencies between f_0 and f_3 and a phase characteristic given by

$$\varphi(\omega) = m\omega - a_0,$$

the use of the above equation permits a graphic determination of the response of the network to a pulse of T duration. The impulse will be regarded as being the superposition of two unit-step functions having opposite signs and being out of phase with each other by time T.

Curves for the response signal for two values of a_0 are given in Figure 17. It will be noted that the signal distortion varies greatly between the two cases and curve *2* for $a_0 = (2K+1)\pi/2$ gives a signal that can hardly be detected in practice.

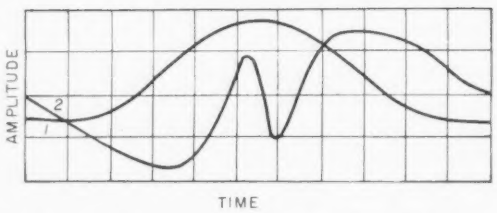


Figure 17—Response to a rectangular pulse of T duration of a circuit having a frequency of $f_0 < f < f_3$ and a phase characteristic $\varphi(\omega) = m\omega - a_0$. With K = any integer, curve *1* is for $a_0 = K\pi$ and curve *2* is for $a_0 = (2K+1)\pi/2$.

7.1.4 Phase Characteristic

$$\varphi(\omega) = m\omega - a_0 + u \sin F$$

The phase characteristic is often bent in either direction as shown in Figure 18. Generally, this curve may be defined to a sufficient approximation by

$$\varphi(\omega) = m\omega - a_0 + u \sin F$$

where

$$F = \pi \frac{\omega - \omega_1}{\omega_1 - \omega_2}.$$

The behavior of a circuit having such a phase characteristic and an amplitude characteristic of $A(\omega) = 1$ for all frequencies between ω_1 and ω_2 is given by

$$S(t) = \frac{1}{\pi}\int_{\omega_1}^{\omega_2} \frac{\sin [\omega(t-m) + a_0 - u \sin F]}{\omega}\, d\omega$$

say

$$S(t) = \frac{1}{\pi}\int_{\omega_1}^{\omega_2} \frac{\sin [\omega(t-m) + a_0]\cos (u \sin F) - \cos [\omega(t-m) + a_0]\sin (u \sin F)}{\omega}\, d\omega.$$

By expanding $\cos (u \sin F)$ and $\sin (u \sin F)$, it becomes[3]

$$\cos (u \sin F) = J_0(u) + 2\sum_1^{\infty} J_{2n}(u) \cos 2nF$$
$$\sin (u \sin F) = 2\sum_1^{\infty} J_{2n-1}(u) \sin (2n-1)F$$

where J_0, J_{2n}, and J_{2n-1} are Bessel functions.

Generally, it is possible to choose m, ω_1, and ω_2 so that u is not important, permitting the above expressions to be limited to the first terms, that is,

[3] Reference Data for Radio Engineers, fourth edition, International Telephone and Telegraph Corporation, New York, 1956; page 1065.

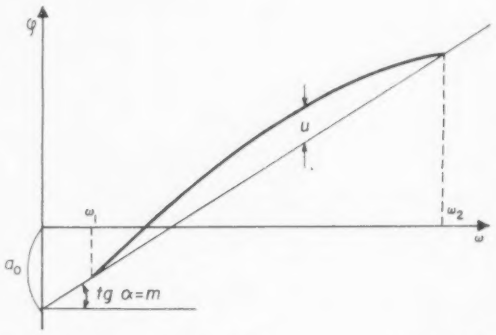


Figure 18—Phase characteristic $\varphi(\omega) = m\omega - a_0 + u \sin F$.

$$\cos(u \sin F) = J_0(u) + 2J_2(u) \cos 2F \qquad \text{and} \qquad \sin(u \sin F) = 2J_1(u) \sin F.$$

The response signal thus becomes

$$S(t) = \frac{J_0(u)}{\pi} \int_{\omega_1}^{\omega_2} \frac{\sin[\omega(t-m)+a_0]}{\omega} d\omega + \frac{2J_2(u)}{\pi} \int_{\omega_1}^{\omega_2} \frac{\sin[\omega(t-m)+a_0]\cos 2F}{\omega} d\omega - \frac{2J_1(u)}{\pi} \int_{\omega_1}^{\omega_2} \frac{\cos[\omega(t-m)+a_0]\sin F}{\omega} d\omega.$$

It is sufficient in each case to express F as a function of ω and expand the different terms to separate the sine-integral and cosine-integral functions. By setting

$$F = k\omega - c$$

say

$$k = \frac{\pi}{\omega_1 - \omega_2} \qquad \text{and} \qquad c = \frac{\pi\omega_1}{\omega_1 - \omega_2}$$

the response signal becomes

$$S(t) = \frac{J_0(u)}{\pi} \int_{\omega_1}^{\omega_2} \frac{\sin[\omega(t-m)+a_0]}{\omega} d\omega + \frac{2J_2(u)}{\pi} \int_{\omega_1}^{\omega_2} \frac{\sin[\omega(t-m)+a_0]\cos(2k\omega-2c)}{\omega} d\omega - \frac{2J_1(u)}{\pi} \int_{\omega_1}^{\omega_2} \frac{\cos[\omega(t-m)+a_0]\sin(k\omega-c)}{\omega} d\omega.$$

After expansion and integration

$$\begin{aligned} S(t) = {} & \frac{J_0(u)}{\pi} \{\cos a_0\, Si\,\omega(t-m) + \sin a_0\, Ci\,\omega(t-m)\}_{\omega_1}^{\omega_2} \\ & + \frac{J_1(u)}{\pi} \{\cos(a_0+c)\, Si\,\omega(t-m-k) + \sin(a_0+c)\, Ci\,\omega(t-m-k) \\ & \quad - \cos(a_0-c)\, Si\,\omega(t-m+k) - \sin(a_0-c)\, Ci\,\omega(t-m+k)\}_{\omega_1}^{\omega_2} \\ & - \frac{J_2(u)}{\pi} \{\cos(a_0-2c)\, Si\,\omega(t-m+2k) + \sin(a_0-2c)\, Ci\,\omega(t-m+2k) \\ & \quad + \cos(a_0+2c)\, Si\,\omega(t-m-2k) + \sin(a_0+2c)\, Ci\,\omega(t-m-2k)\}_{\omega_1}^{\omega_2}. \end{aligned}$$

The first term represents the behavior of a circuit having a phase characteristic represented by a straight line crossing a_0, having a slope m, the amplitude of which is reduced as the ratio $[J_0(u)]/\pi$.

The other two terms correspond to the behavior of circuits having linear phase characteristics but having different slopes and different origins.

The bending of the phase characteristic $\varphi(\omega)$ thus causes spreading of the transient phenomena.

7.1.5 Application to Loaded Cable

To determine graphically the behavior of a circuit in the loaded *H88* cable of 264-kilometer (164-mile) length to a unit-step transition, the frequency range from f_0 to f_3 may be divided into three parts designated f_0–f_1, f_1–f_2, and f_2–f_3. The response over each part of the frequency range may then be defined. The total response will be the sum of the partial responses.

7.1.5.1 Response for f_0–f_1

In the lowest selected band of f_0–f_1, let $f_0 = 100$ and $f_1 = 250$ cycles per second. The response may be calculated through use of the procedure

of section 7.1.4 by employing the following numerical values taken from Figure 12.

$a_0 = 6.5\pi$, $m = 22.7 \times 10^{-3}$, and $u = 0.4\pi$.

7.1.5.2 Response for f_1–f_2

For the next frequency band f_1–f_2, let $f_1 = 250$ and $f_2 = 2000$ cycles per second. The phase characteristic approaches the straight line *CD* shown in Figure 11 from which $a_0 = 1.7$ and $m = 13 \times 10^{-3}$. The response will be obtained using the method outlined in section 7.1.3.

7.1.5.3 Response for f_2–f_3

For the upper band f_2–f_3, let $f_2 = 2000$ and $f_3 = 3000$ cycles per second. Calculations are made as in section 7.1.4 using $a_0 = 12.5\pi$, $m = 15.75 \times 10^{-3}$, and $u = -0.6\pi$ taken from Figure 16.

7.1.5.4 Response for f_0–f_3

The total response to a unit-step transition is obtained by adding graphically the partial responses obtained for the three frequency bands.

To deduce the behavior for a code group including several pulses, each pulse may be regarded as made up of two unit-step transitions of opposite signs separated in time by the duration of the pulse. The results of the application of this method to the transmission of the code group *000110100* at a speed of 2000 bauds is given in Figure 19.

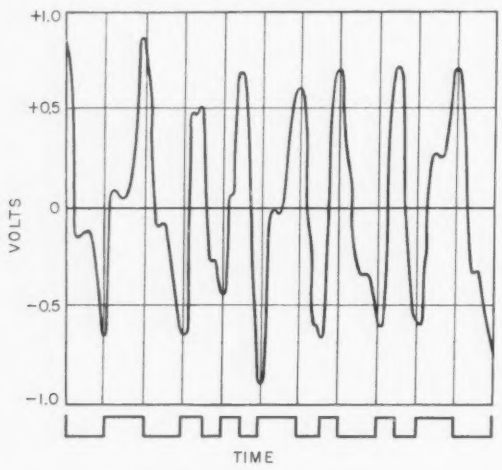


Figure 19—Graphic determination of the response to the signal for the coded information of *00011010100* transmitted at 2000 bauds.

There is substantial resemblance between this calculated waveform and that of an oscillogram given in Figure 20.

Maximum distortion occurs at the transition between long and short signals. The half-period

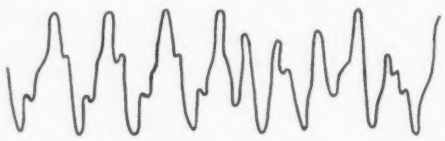


Figure 20—Oscillogram of response to code group *000000001101010100* transmitted at 2000 bauds.

of a signal *1* is therefore lengthened or shortened as a result of one or more cycles of high-frequency signal appearing at the pulse edge.

If the binary signals are continuously *0* or *1*, this spurious effect occurs at the same place for all digits, producing a waveform that is symmetrical about the zero level.

If the code changes from one binary character to the other, this spurious signal is shifted markedly and produces a sharp distortion.

As transmission speed increases, the time between successive transitions decreases. At certain speeds, the transients resulting from phase distortion of the higher frequencies become superimposed in part on each other. The spurious signals at the edges of the pulses then produce more-damaging distortion.

As transmission speed decreases, the time interval between transitions increases and the transient caused by phase distortion at the lower frequencies produce a ripple waveform at the receiving end.

The above statements show clearly the importance of phase distortion and its production of transients. This distortion is generated not only by the line but by repeaters, carrier equipment, filters, and other elements that may be inserted in the transmission system. It is advisable, therefore, to examine all such elements to minimize their contributions to distortion.

The distortion discussed above occurred with unmodulated signals. It will be the same for modulated signals by replacing, at the frequency axis origin, the 0 point by the carrier frequency.

8. *Tests on Unloaded Cable*

Tests were made on an artificial circuit corresponding to an unloaded cable. The line constants were

Resistance = 54.5 ohms per kilometer
= 87.7 ohms per mile

Capacitance = 33.6 nanofarads per kilometer
= 54.1 nanofarads per mile.

These values are approximately those of subscriber cable of 0.9-millimeter diameter, which is approximately 19 American Wire Gage.

The attenuation and phase-shift characteristics of a 26-kilometer (16.2-mile) length of cable are shown in curves *1* of Figures 21 and 22. These distortions may be compensated for by equalizers. Curves *2* of these figures show how the equalizer produces a nearly constant attenuation and a phase shift that is approximately linear with frequency over the required frequency range.

Measurements were made for nearly the same conditions outlined in section 6. The insertion of an equalizer to compensate for distortion increased the attenuation. A 2-wire repeater providing a gain of 4.8 nepers (41.8 decibels) over the desired frequency range was employed to make up for the increased attenuation. Table 4 gives the results of the tests.

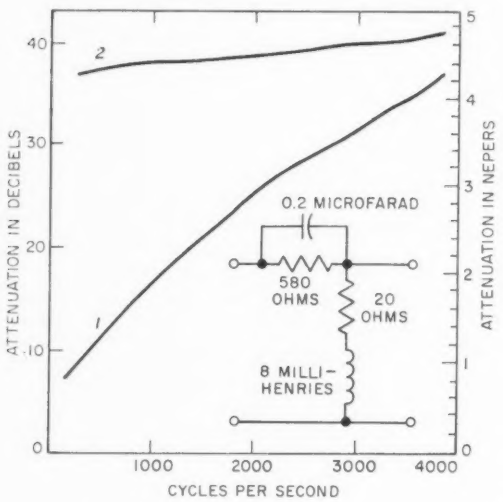


Figure 21—Attenuation is plotted against frequency for a 26-kilometer (16.2-mile) unloaded cable in curve *1*. Curve *2* is with the equalizer shown in the diagram.

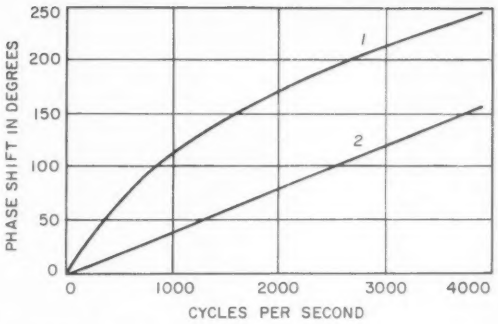


Figure 22—Phase shift is plotted against frequency for the unloaded cable in curve *1* and in curve *2* with the equalizer shown in Figure 21.

TABLE 4

TESTS ON UNLOADED CABLE

Transmission Type	Cable Length		Transmission Speed in Bauds	Operating Margin	
	Kilometers	Miles		Nepers	Decibels
Unmodulated	26	16	2000	5.35	46.5
Unmodulated	26	16	4000	5.20	45.2
Unmodulated	52	32	4000	2.00	17.4
Modulated*	26	16	2000	3.45	30.0

* Carrier frequency = 2000 cycles per second.

Although the test equipment was not designed for transmission speeds above 4000 bauds, it is unlikely that the unloaded cable would be limited to this speed.

9. *Noise*

The permissible noise limits have been obtained by superimposing the output of a white-noise generator on the received signal, after transmission of the signal without distortion through an attenuating network.

For unmodulated transmission, the limiting signal-to-noise ratio expressed in peak-to-peak values varies from 0.7 neper (6 decibels) for a transmission speed of 500 bauds to 1.4 nepers (12 decibels) at 3000 bauds. For modulated transmission this limit is 1.4 nepers (12 decibels) regardless of the speed.

10. Duty Factor

It is interesting to compare the transmission speeds for the preceding tests with the capacity of the channels to determine the duty factor.

$$\text{Duty Factor} = \frac{\text{Transmission Speed}}{\text{Channel Capacity}}.$$

According to information theory

$$C = 2F \log_2 (S/B)$$

where

C = transmission capacity in bits
F = frequency range in cycles per second
S = signal level
B = noise level.

In practice, for a binary system in which $S/B = 2$, the transmission capacity is

$$C = 2F.$$

A comparison of various duty factors with that for carrier telegraphy is given in Table 5.

It will be noticed that for a given cable and a constant frequency bandwidth, the transmission speed decreases as the cable length increases. This is contrary to information theory, which defines channel capacity as

$$C = 2F \log_2 (S/B).$$

The discrepancy results from the fact that the equation neglects the effects of phase distortion.

The transmission speed for modulated signals will be half that for unmodulated signals, for the modulated system requires twice the bandwidth of the unmodulated transmission.

11. Conclusions

The experimental results on lines of varying length and characteristics show that the equipment can transmit numerical data at high speed over telephone channels. The maximum transmission speed depends not only on the available frequency bandwidth but also on the phase-characteristic distortion.

In the case of leased (unswitched) circuits of average length, a speed of 2000 to 3000 bauds can be obtained over cable of the *H88* type and of 3000 to 4000 bauds over unloaded cable. The limit for unloaded cables has not been clearly established but should exceed the value given.

For switched circuits, the channel characteristics and noise spectrum make amplitude or frequency modulation necessary and the transmission speed will be reduced to about 750 bauds.

12. Acknowledgment

This work was greatly facilitated by the cooperation of the Société Nationale des Chemins de fer Français, who made available the loaded cables intended for the remote-control system much in advance of their being put in regular service.

TABLE 5

COMPARISON OF DUTY FACTORS

System	Transmission Circuit	Speed in Bauds	Frequency Limits in Cycles per Second	Real Transmission Capacity of Channel in Bits per Second	Duty Factor of Channel
High-Speed Wide-Band Transmission	Unloaded Cable	4000	100–4000	7800	0.51
	H88 Cable of 100 Kilometers (62 miles)	3000	100–3500	6800	0.44
	H88 Cable of 264 Kilometers (164 Miles)	2000	100–3400	6600	0.30
		700	100–3400	6600	0.11
Voice-Frequency Telegraph	1 Voice-Frequency Channel	50	120	240	0.21

Some Performance Parameters of Silicon Junction Power Rectifiers*

By D. R. COLEMAN

Standard Telephones and Cables Limited; London, England

ELECTRICAL performance of silicon power rectifiers—seen in terms of the power output of a conversion connection—is related to their characteristics and is shown to be dependent on the junction temperature. Temperature control is discussed from the thermal-resistance standpoint and a method is given for the calculation of power dissipation and connection ratings for silicon rectifiers. Mention is made of some considerations necessary for satisfactory transient performance.

1. P-N Junction

1.1 Electrical Characteristics and Temperature Dependence

The electrical characteristics of silicon junction rectifiers in the forward and reverse directions are well known[1–4]. The performance of such rectifiers in power conversion connections is directly related to those characteristics, and in this sense the variations of characteristic with junction temperature are of critical importance.

In general terms, the forward characteristics suggest an improved performance with increase of junction temperature, since a given current in the working region produces a decreasing voltage drop with increasing temperature. On the other hand, the reverse characteristics indicate a degradation in performance, in that at a fixed working voltage the current increases with rise of junction temperature.

A typical temperature coefficient of forward voltage drop would be $-0{\cdot}1$ per cent per degree centigrade, showing the effect on performance to be slight. The reverse current, however, may vary considerably. Although the saturation current can increase by as much as 15 per cent per degree centigrade,[2,5] the total cell current variation with temperature is usually less by virtue of the less-sensitive leakage currents present. The overall effect on performance is not necessarily as severe as these variations might suggest, since the effective rectification ratio of a silicon rectifier is usually greater than 10^5.

1.2 Junction Losses

It is obvious that operation of a junction rectifier in a conversion connection will produce electrical losses, which may be assessed by the methods described in section 3. The consequent heat generation from those losses means that the rectifier junction will always operate at a temperature greater than its surroundings.

1.3 Junction Temperature

Apart from its influence on the electrical characteristics of the rectifier, operation at an extremely high junction temperature may produce undesirable effects in the constituents of the rectifier assembly. Improvements in processing and encapsulation have been shown to give rectifiers capable of operation at a junction temperature of 150 degrees centigrade for very-long periods; there is reason to believe, however, that a shorter life results from continuous operation at, for example, a junction temperature of 200 degrees centigrade.

For some applications a short life is acceptable, and in these cases an uprating of operating temperature can give useful increases in the

* Reprinted from *Electronic Engineering*, volume 32, pages 38–102; February, 1960.

[1] E. F. Losco, "Properties of Silicon Power Rectifiers," *Transactions of the American Institute of Electrical Engineers*, part 1, volume 74, and *Communication and Electronics*, number 17, pages 106–111; March, 1955.

[2] G. Finn and R. Parsons, "Some Basic Physical Properties of Silicon and How they Relate to Rectifier Design and Application," *Institute of Radio Engineers Transactions on Component Parts*, volume CP3, pages 110–113; December, 1956.

[3] H. W. Henkels, "Fused Silicon Rectifier," *Transactions of the American Institute of Electrical Engineers*, part 1, volume 75, and *Communication and Electronics*, number 28, pages 733–746; January, 1957.

[4] F. W. Gutzwiller, "Rating and Application of Germanium and Silicon Rectifiers," *Transactions of the American Institute of Electrical Engineers*, part 1, volume 75, and *Communication and Electronics*, number 28, pages 753–757; January 1957.

[5] J. Millman, "Vacuum Tube and Semiconductor Electronics," McGraw-Hill Book Company, New York, New York; 1958.

electrical ratings obtained. For a general-purpose rectifier a continuous operating temperature consistent with long life has the additional advantage of permitting a greater momentary temperature to be sustained if necessary, under transient conditions of forward voltage or reverse current surges.

2. *Thermal Considerations*

2.1 Junction Losses and Temperature Difference

Each watt of junction dissipation causes a heat-flow rate of approximately 0·24 gramme-calories per second from the junction, that heat flow producing a temperature difference across any thermal resistance in its path, such that under steady-state conditions,

$$T = W\theta, \text{ degrees centigrade}$$

where W is the heat flow rate in watts and θ is a thermal resistance in degrees centigrade per watt. The operating junction temperature is therefore greater than the cooling-medium temperature, since a thermal resistance cannot be avoided in the extraction of the heat produced.

2.2 Types of Rectifier

For thermal considerations of rectifier performance, there are two separate types of rectifier: (A) those intended for convection cooling and (B) those intended for conduction cooling.

The first class includes wire-ended and integral-fin rectifiers, where the rectifier dissipating properties in a given medium are unalterable. The second class describes the stud-ended and similar rectifiers, whose dissipating properties may be varied by the attachment of different fins or heat sinks. The differences between the classes may be readily understood by using electrical analogues of their thermal circuits.

2.3 Wire-Ended Rectifiers

With a wire-ended or integral-fin rectifier, continuous heat flow is from the junction to the case, and thence by combined convection and radiation, and conduction through the terminations, to the cooling medium (as shown in Figure 1).

The internal temperature gradient is then

$$T_{jc} = W\theta_{ja}$$

and will not be affected greatly by changes in the ambient environment.

The external temperature gradient is

$$T_{ca} = W\theta_d$$

and will be significantly altered by the effectiveness of the cooling medium. The value of θ_d for

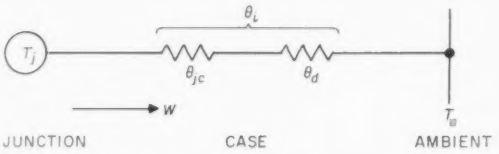


Figure 1—Heat generation and thermal-resistance path of a wire-ended rectifier. θ_{jc} = internal thermal resistance, θ_d = device effective thermal resistance, and θ_i = intrinsic thermal resistance.

an air-cooled rectifier of 1-watt rated dissipation is reduced typically to 1/6th by immersion in oil, to give one illustration.

The overall thermal resistance of the rectifier is the sum of the internal and external components

$$\theta_{ja} = \theta_{jc} + \theta_d \qquad (1)$$

and is constant for a given type of cooling, for example convection air cooling or forced-air cooling. It may therefore be considered as an intrinsic property of the rectifier in a stated cooling medium, in which event the junction-temperature rise will be directly related to the junction losses:—

$$T_{ja} = W\theta_i, \qquad (2)$$

where θ_i is the intrinsic thermal resistance, equal to θ_{ja} in (1).

The intrinsic thermal resistance is directly measurable, by continuous dissipation of a known junction loss and the detection of the steady junction temperature by reference to its temperature-dependant forward characteristic[6]. Temperature sensing by use of the reverse characteristic is also possible[7].

[6] F. Spitzer and B. Steinhauser, "Temperaturmessungen an Silizium gleichrichtern," Congrès International de Bruxelles; 1958.

[7] C. D. Simmonds, Semiconductors Ltd., Applications Report *307A*.

2.4 Stud-Mounted Rectifiers

Under steady-state operating conditions, the heat generated at the junction in a stud-ended rectifier may be considered as flowing through an internal thermal resistance and then through an external thermal resistance to the cooling medium (as shown in Figure 2).

The internal thermal resistance from the junction to the stud may be assumed constant. The external thermal resistance is conveniently thought of as a parallel circuit of a constant thermal resistance of the rectifier alone, shunted by a thermal resistance due to the heat sink. The value of the heat-sink thermal resistance will be variable.

Without an added heat sink the considerations of section 2.3 apply, and the overall thermal resistance (the rectifier intrinsic thermal resistance) is

$$\theta_i = \theta_{js} + \theta_d, \tag{3}$$

in which case θ_d varies according to the cooling medium (or ambient) conditions, but θ_{js} remains virtually constant. For a given maximum junction temperature T_{jmax}, the greatest dissipation possible without a heat sink will be

$$W_{imax} = \frac{T_{jmax} - T_a}{\theta_i} \tag{4}$$

and is called the maximum intrinsic dissipation at ambient temperature T_a.

If the rectifier is attached to a heat sink or fin of thermal resistance θ_f, the thermal resistance from the stud to the cooling medium will be

$$\theta_{sa} = \frac{\theta_f \theta_d}{\theta_f + \theta_d}. \tag{5}$$

Thus the total thermal resistance to heat flow from the junction becomes

$$\theta_{ja} = \theta_{js} + \theta_{sa}. \tag{6}$$

A given fin of thermal resistance θ_f may have a different effect with different rectifiers (due to variations of θ_d between types of rectifier) and will have a considerably different effect in different cooling media since both θ_f and θ_d are dependent on the cooling medium used.

In the case of a zero thermal resistance (infinite heat sink) being attached to the stud, that is, $\theta_f = 0$, the external thermal resistance becomes zero (from (5)) but the temperature drop from the junction to the stud is unaltered at

$$T_{js} = W\theta_{js} \tag{7}$$

and the stud temperature is $T_s = T_j - W\theta_{js}$. If the rectifier is now operated at its maximum dissipation W_{max}, the maximum junction temperature is produced at a stud temperature of

$$T_{smax} = T_{jmax} - W_{max}\theta_{js}. \tag{8}$$

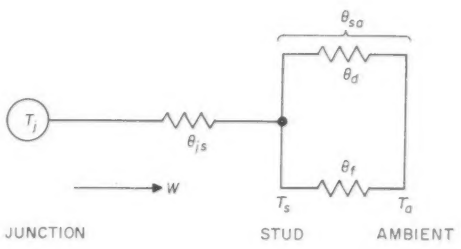


Figure 2—Heat generation and thermal-resistance paths of a stud-mounted rectifier. θ_{js} = internal thermal resistance, θ_d = device effective thermal resistance, and θ_f = heat-sink thermal resistance.

W_{max} is thus the ultimate dissipation obtainable at stud temperatures up to T_{smax}.

The performance of a stud-ended rectifier can therefore be determined in terms of a given stud temperature or of a given heat-sink thermal resistance and cooling medium.

2.5 Heat Sinks and Cooling Methods

Separately attached heat sinks for silicon power rectifiers intended for natural air convection, forced air draught, and liquid coolant immersion may be of the single or multi-fin types; much ingenuity has been shown in the development of special forms and sections to permit ready interlocking and formation of bus connections.

Their thermal resistance is measurable, as indicated in section 2.4, or may be calculated by a variety of methods[8,9]. The θ_f values can vary from, for example, 14 degrees centigrade per watt for a simple 3-by-3-inch (7·6-by-7·6-

[8] W. Luft, "Design of Fins for Cooling Semiconductors," *Electrical Manufacturing*, volume 60, pages 98–103; November, 1957.

[9] H. L. Aronson, "Applying Power Transistors to Control," *Control Engineering*, pages 78–87; October, 1956.

centimetre) fin of 1/16-inch (0·16-centimetre) copper to 0·4 degree centigrade per watt for a 7-inch (17·8-centimetre) length of a multi-finned 6-by-4-inch (15·2-by-10·2-centimetre) extruded aluminium section, suitable for 1-watt- and 150-watt-dissipation rectifiers respectively under natural air cooling conditions.

Higher dissipations, or the need to reduce the temperature rise, demand more effective cooling methods such as heat extraction by a circulating liquid coolant system. A heat exchanger of this form, for example, can dissipate 1·2 kilowatts with an oil temperature rise of 5 degrees centigrade, by an oil flow of 2 gallons (9·1 litres) per minute through a section of area 1·75 square inches (11·3 square centimetres), over a 5-foot (152-centimetre) length of hollow aluminium bar.

3. Steady-State Conditions

3.1 Determination of Forward Losses

The forward characteristic of a silicon rectifier may reasonably be idealized, as shown in Figure 3, to give the relationship

$$V = a + bI$$

where a and b are constants.

Constant a is termed the threshold voltage (volts) and b is the slope resistance (ohms).

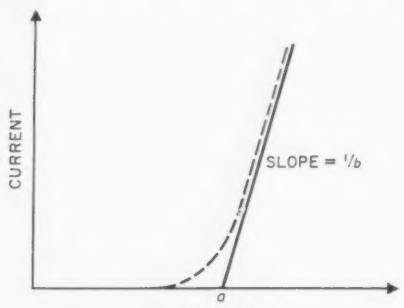


Figure 3—Electrical characteristics in the forward direction. The idealized characteristic is the solid line and the broken line is the typical characteristic obtained in practice.

Practical values are found by extrapolating the line joining the voltage values at T_{jmax} corresponding to $\bar{I}$ and $\hat{I}$, where $\bar{I}$ is the mean current, and $\hat{I}$ the peak current carried in a rectifier connection. The slope of that line is equal to $1/b$, and its intercept with the voltage axis gives the threshold voltage.

When conducting in the forward direction, the instantaneous voltage drop will be

$$v = a + bi$$

where i is the instantaneous current. The instantaneous power dissipation is given by

$$vi = ai + bi^2, \text{ watts.} \qquad (9)$$

With a sinusoidal voltage supply and a resistive load, the forward current through the rectifier may be assumed to be a part-sinusoid symmetrical about a peak value $\hat{I}$, and flowing for a conduction period 2θ radians, where the value of 2θ depends on the rectifier connection. For example, $2\theta = \pi$ radians in a single-phase connection, or $2\theta = \pi/3$ radians in a three-phase push-pull connection.

TABLE 1

Rectifier Connection	Forward Loss per Rectifier W_f Watts	
	Peak Current $\hat{I}$	Rectifier Mean Current $\bar{I}$
Single-Phase Half-Wave / Push-Pull / Bridge	$0{\cdot}32a\hat{I} + 0{\cdot}25b\hat{I}^2$	$a\bar{I} + 2{\cdot}46b\bar{I}^2$
Three-Phase		
Half-Wave	$0{\cdot}28a\hat{I} + 0{\cdot}24b\hat{I}^2$	$a\bar{I} + 3{\cdot}10b\bar{I}^2$
Push-Pull	$0{\cdot}16a\hat{I} + 0{\cdot}15b\hat{I}^2$	$a\bar{I} + 6{\cdot}01b\bar{I}^2$
Bridge	$0{\cdot}32a\hat{I} + 0{\cdot}30b\hat{I}^2$	$a\bar{I} + 3{\cdot}00b\bar{I}^2$

Equation (9) may then be rewritten

$$vi = a\hat{I}\cos\theta + b\hat{I}^2\cos^2\theta$$

and the mean power dissipation over the conduction period derived as

$$\overline{vi} = \frac{a\hat{I}}{2\theta}\int_{-\theta}^{+\theta}\cos\theta\,d\theta + \frac{b\hat{I}^2}{2\theta}\int_{-\theta}^{+\theta}\cos^2\theta\,d\theta. \qquad (10)$$

Since there is a fixed relationship (with a given rectifier connection) between the peak current $\hat{I}$ and the mean current $\bar{I}$ per rectifier, and because the conduction period is repeated at supply frequency, the mean power dissipation per cycle $\overline{VI}$ may be expressed in terms of a and b. $\overline{VI}$ is in fact the forward loss per rectifier or W_f, and is given in Table 1.

The same principle of solving the integral $\int vi\,dt$ to determine the total forward loss per

cycle may be employed for other current waveforms. For example, square-wave conduction in the bridge connections may be compared with the part-sinusoid condition as shown in Table 2.

TABLE 2

Rectifier Connection	Forward Loss per Rectifier W_f	
	Square-Wave	Part-Sinusoid
Single-Phase Bridge		
$\hat{I}$	$0{\cdot}5a\hat{I}+0{\cdot}5b\hat{I}^2$	$0{\cdot}32a\hat{I}+0{\cdot}25b\hat{I}^2$
$\bar{I}$	$a\bar{I}+2b\bar{I}^2$	$a\bar{I}+2{\cdot}46\bar{I}^2$
Three-Phase Bridge		
$\hat{I}$	$0{\cdot}33a\hat{I}+0{\cdot}33b\hat{I}^2$	$0{\cdot}32a\hat{I}+0{\cdot}30b\hat{I}^2$
$\bar{I}$	$a\bar{I}+3b\bar{I}^2$	$a\bar{I}+3b\bar{I}^2$

3.2 Determination of Reverse Losses

The total loss in the reverse direction is found from $\int vi\,dt$, integrated over the complete cycle. The voltage waveform is largely dependent on the rectifier connection being used, and the reverse current varies according to the particular rectifier characteristic. It is therefore difficult to generalize, but since the reverse losses in silicon rectifiers form a relatively small proportion of the total dissipation, it is usually sufficient to approximate their value and consider this as a constant in normal use.

If the reverse voltage is taken to be at its crest value for a duration appropriate to the connection (for example, for 0·67 of the cycle in a 3-phase bridge) and the reverse current is of the maximum value at the maximum rated junction temperature, the reverse loss thus found

$$W_r = \tfrac{2}{3} V_{cw} I_r, \text{ watts,}$$

(where V_{cw} is the rated crest working voltage), is the highest value likely in practice, and may be taken as constant without detracting greatly from the calculated performance of the rectifier under other circumstances.

3.2.1 *Thermal Instability*

It must be remembered that the rectifier characteristics are temperature dependent. The forward characteristic of a silicon *pn* junction has a negative temperature coefficient; at a given forward current the forward losses decrease with increase of junction temperature. This tends to a temperature stabilizing situation, but the coefficient is relatively small.

The reverse current, however, increases with junction temperature at voltages in the working region. The reverse losses due to these voltages and currents may be expressed in the form[10]

$$W_r = W_a \exp(cT) \tag{11}$$

where

W_r = loss at junction temperature of T_j degrees centigrade

W_a = loss at junction temperature of T_a degrees centigrade

c = constant of temperature variation of I_r

$T = (T_j - T_a)$ in degrees centigrade.

Now the total power loss

$$W = W_r + W_f \tag{12}$$

and the heat-flow rate

$$W_d = T/\theta \tag{13}$$

where

W_d = power dissipation to ambient

θ = thermal resistance in degrees centigrade per watt from junction to ambient.

Substituting (11) in (12),

$$W = W_a \exp(cT) + W_f. \tag{14}$$

Equation (14) represents the power causing heat generation, in which W_f may be assumed constant. Equation (13) represents the power-dissipating properties of the rectifier.

Thermal instability will occur if the rate of generation of heat exceeds the rate of dissipation of heat; conversely stability is assured if the dissipation rate exceeds the generation rate. If the two rates are equal, the maximum possible heat flow for thermal stability is present.

This situation is found by equating the derivatives of (13) and (14), that is,

$$1/\theta = cW_a \exp(cT)$$

or

$$\exp(cT) = 1/cW_a\theta. \tag{15}$$

[10] D. K. Bisson, "Rating and Application of a Silicon Power-Rectifier," American Institute of Electrical Engineers' Publication *93*; June, 1957.

Inserting (15) in (14),

$$W_{max} = W_f + W_a(1/cW_a\theta)$$

or since W_f is assumed constant,

$$W_{rmax} = 1/c\theta. \qquad (16)$$

The factor c is deduced from the rectifier reverse characteristics; its maximum theoretical value for the true saturation current is approximately 0·115, any surface leakage currents effectively reducing this value. The factor θ can be reduced in the limiting case to θ_{js} for a stud-mounted rectifier; the maximum, on the other hand, will be θ_i $(= \theta_{js} + \theta_d)$ with this type of rectifier, or θ_i $(= \theta_{ja})$ for a wire-ended design.

The theoretically maximum reverse loss for thermal stability will thus be, in the worst operating condition,

$$W_{rmax} = 1/c\theta_i$$
$$= 8{\cdot}7/\theta_i, \text{ watts.} \qquad (17)$$

This clearly may be exceeded for more-efficiently cooled rectifiers, and may also be increased in the practical case by the existence of surface leakage currents.

3.3 Division of Losses

Section 2 discussed the relation between junction temperature and junction dissipation in a rectifier under given cooling conditions. Sections 3.1 and 3.2 show how the losses caused by forward conduction and reverse blocking may be evaluated.

To assess the possible performance of a rectifier when used in a conversion connection, a division of the dissipation into forward and reverse losses must be made. The rectifier could perform equally well as a blocking rectifier (100-per-cent reverse losses) or as a device conducting continuously in the forward direction (zero reverse losses). Intermediately a high forward current and low reverse voltage could result in the same total dissipation as the converse in a rectifier connection.

3.3.1 Assignment of Ratings

For general-purpose power-rectifier applications, especially with a range of rectifiers of one case design, the crest-working-voltage rating assigned to a given rectifier will be controlled either by the maximum reverse dissipation allowable for thermal stability, or by the breakdown (turnover) characteristic of the junction, in which region the rectifier should not normally be operated. If neither of these considerations is overriding, it is probable that the application itself will impose a desirable limit on the reverse current, and consequently the reverse dissipation. In general the reverse loss should not normally exceed 10 per cent of the total dissipation.

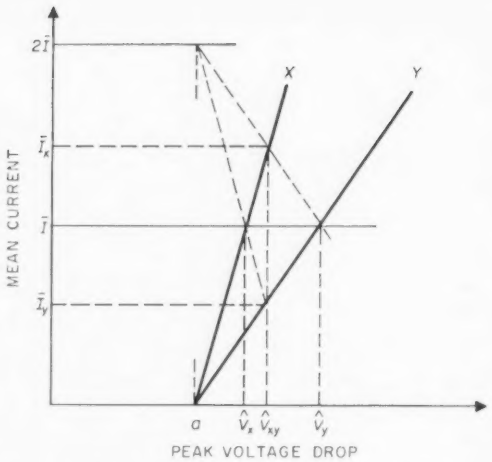


Figure 4—Parallel operation of rectifiers X and Y causes the total mean current $2\bar{I}$ to be shared as $\bar{I}_x$ and $\bar{I}_y$.

The remainder can in principle be produced by the forward losses, equivalent therefore to a certain mean-forward-current rating in the particular connection desired. As Table 1 shows, the same forward dissipation may be caused by different rated currents in different connections. The relationships between the input voltage equivalent to a required crest working voltage and the resulting output voltage are well known for the more-common connections; thus these ratings can be calculated for the connection desired.

Assuming the reverse loss and the rated crest working voltage to remain constant, current ratings may be deduced in this way for all variants of the cooling method applied to the rectifier, the ambient conditions in which the rectifier operates, or the type of heat sink used.

3.4 Parallel Operation

When rectifiers are operated in parallel in a conversion connection, the voltage across each rectifier will be the same at any instant. During the period of reverse stress, the reverse losses in one rectifier will not be affected by the presence of another, since the applied reverse voltage is a circuit property only.

During the period of forward conduction, however, the common voltage across paralleled rectifiers may cause quite different currents to flow in each rectifier, due to their dissimilar forward characteristics. As a result, the maximum rated dissipation of one of the rectifiers may be exceeded.

In the absence of any current-balancing influence (from series equalizing resistors[2] or magnetically coupled series reactors) excess dissipation can be avoided by limiting the dissimilarity of characteristics of paralleled rectifiers.

Consider one rectifier X carrying a mean forward current $\bar{I}$ in a 3-phase bridge connection. The forward losses will be

$$W_f = \bar{I}\hat{V}, \text{ watts,} \qquad (18)$$

$\hat{V}$ being the forward voltage at $\hat{I} = 3\bar{I}$. If $\bar{I}$ is the rated mean current for that rectifier, we may call its peak voltage drop $\hat{V}_x$ (that is, at $3\bar{I}$).

The parallel connection of a second rectifier Y, of peak voltage drop $\hat{V}_y$, will cause a common peak voltage drop $\hat{V}_{xy}$ when a total mean current of $2\bar{I}$ is taken from the two. Figure 4 shows that the total current $2\bar{I}$ is composed of $\bar{I}_x$ from rectifier X and $\bar{I}_y$ from rectifier Y. The increase in mean current through the first rectifier from $\bar{I}$ to $\bar{I}_x$ produces an increase in forward loss of

$$\delta W = (\bar{I}_x\hat{V}_{xy}) - (\bar{I}\hat{V}_x), \text{ watts,} \qquad (19)$$

which is controlled primarily by the relation between the peak voltage drop of the two rectifiers, $\hat{V}_x$ and $\hat{V}_y$ (see Figure 5).

The increase in dissipation must not allow rectifier X to exceed its rated maximum dissipation if safe operation is to be assured. The greatest value of δW possible is dependent on the nominal loss $\bar{I}\hat{V}_x$ of rectifier X; the higher the peak voltage drop $\hat{V}_x$, the less will be the allowable increase δW. Thus to Figure 5 may be added a curve showing the maximum allowable increase in forward loss, from zero at the voltage corresponding to the rated maximum forward dissipation, to that dissipation at zero voltage.

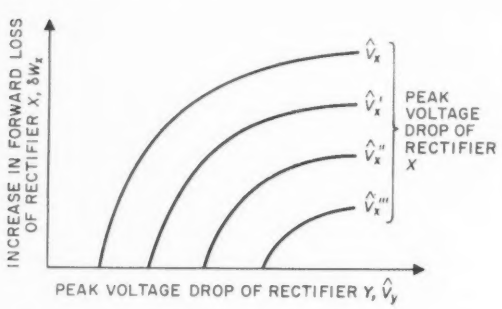


Figure 5—Paralleling rectifier X with rectifier Y at double the rated current of each causes an increase δW_x in the dissipation of rectifier X.

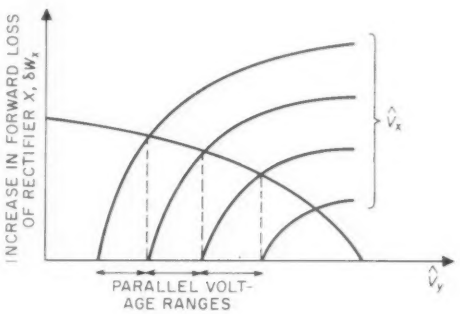


Figure 6—The maximum forward loss-increase curve determines the forward voltage range of rectifiers that may be paralleled.

The intersection of this maximum forward loss-increase curve with the $\hat{V}_x$ curve gives the limit peak voltage drop $\hat{V}_y$ of a rectifier that may be safely operated in parallel. Continuation of this construction evaluates the voltage range of rectifiers that may be paralleled in this way (see Figure 6).

3.5 Series Operation

Rectifiers operated in series carry the same current at all instants. Forward conduction losses in one rectifier are therefore unaffected by the presence of others in the series if thermally

separate, and each rectifier will perform in the same way as it would if operated separately. In the case of thermal interconnection (see section 3.6) a temperature-equalizing effect will naturally ensue.

The reverse-voltage distribution, on the other hand, will be determined by the reverse characteristics of the individual rectifiers in the series and will not necessarily be equal.

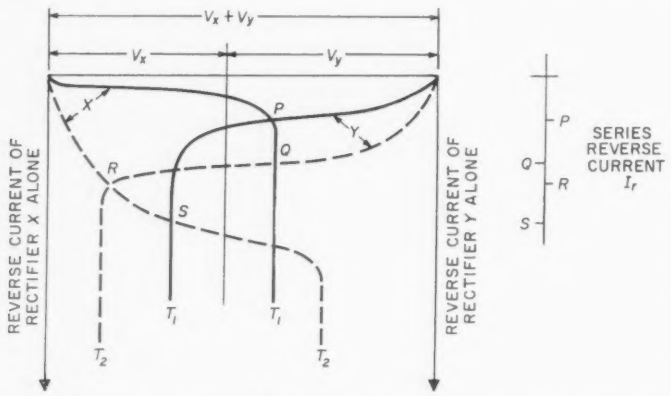


Figure 7—Reverse voltage distribution with two series-connected rectifiers. T_1 and T_2 are junction temperatures with $T_2 > T_1$.

Consider two rectifiers X and Y operated in series, whose rated reverse voltages are V_x and V_y. Operation at an applied voltage equal to $(V_x + V_y)$ will cause a voltage distribution shown by the points P, Q, R, S in Figure 7, the actual operating point being controlled by the relative junction temperatures of the two rectifiers.

If both junctions are at temperature T_1, the operating point is P; it is seen that rectifier X is stressed at greater than its rated voltage V_x. Should this cause an increase in junction temperature, the operating point will tend to move towards S, thus reducing the stress across X. This effect, together with the simultaneous understressing initially of rectifier Y, tends to stability.[4] The diagram also demonstrates the desirability of defining the rated crest working voltage of the rectifiers at their maximum operating junction temperature.

The use of shunt resistance to impose an equalized voltage distribution in the series is not normally desirable, since it reduces the sensitivity of the self-stabilizing effect and could be dangerous in the event of a resistor open-circuit. Shunt resistance and capacitance is sometimes necessary, however, to equalize the distribution of high-frequency transient voltages by swamping the rectifier capacitance and the unequally distributed capacitances to earth of the rectifiers.

3.6 Thermal Interconnection

It will be readily appreciated that stud-ended rectifiers on identical but separate heat sinks may operate at different temperatures when carrying the same forward current, due to differences in the total dissipation of particular examples of the same rectifier type. Especially with the higher-power rectifiers, small differences of forward characteristic can thus cause considerable variations in junction operating temperatures. Mounting on a common heat sink can reduce these temperature differences and assist towards reverse-voltage equalization in series connection.

Less apparent, probably, is the solution to the problem of ensuring acceptable reverse-voltage distribution in an overall encapsulation of series-connected junctions. The encapsulation could, for example, be of the wire-ended form where heat extraction is possible only from the lead terminations, and the separate junctions may have differences in forward characteristic that would cause significant inequalities in operating temperature if the junctions were separately encapsulated and cooled. It can be shown that by manipulation of the internal thermal resistances, a set of asymmetric parallel paths can be presented to heat flow from each junction, and the junction-temperature rises made less susceptible to differences in dissipation of the individual junctions.

4. Transient Conditions

The previous sections refer to the steady-state performance of silicon power rectifiers, where the continuous operating temperature of the junction is the main concern.

Their performance under transient conditions is usually dictated by the two more-important criteria of the effects of the junction thermal time constant and of the minority-carrier storage.

4.1 Junction Thermal Time-Constant

In the same way that a heat sink possesses a thermal capacity that influences the rate of rise of rectifier stud temperature, there is a thermal capacity associated with the internal thermal resistance of the rectifier that will control the junction-temperature rise under transient conditions. The junction thermal time-constant can be evaluated from its cooling properties after pulsed heating[6,7,11].

In practice, the transient heating effects on the junction of a forward-current overload determine on the one hand the type of duty that can be performed by the rectifier from a given steady-state condition, and on the other hand the form and speed of operation of protective devices that are necessary to preserve the rectifier characteristics.

The junction cooling characteristic will to some extent be influenced by the rectifier external thermal resistance, whose properties will thus have a bearing on the intermittent duty possible from the rectifier.

[11] B. Reich, "Thermal Considerations in Power Transistor Applications," *Electrical Manufacturing*, volume 59, pages 162–163, 302, 304; May, 1957.

4.2 Minority-Carrier Storage

The effect of minority carriers present in the base region after commutation is to generate a reverse-voltage transient on being swept out, due to the presence of inductance in the circuit. In polyphase connections, these transients appear in the appropriate phase relationship across each rectifier.[4,12]

In most cases, it is possible by suitable circuit design to reduce the transient voltages to a value consistent with the reverse-voltage rating of the rectifiers, whose connection performance need not then be restricted by their minority-carrier storage properties.

5. Conclusions

The parameters introduced in this paper are pertinent to the assessment of performance of individual rectifiers; the considerations for series and parallel arrangements concern the performance of rectifiers in only one arm of a conversion connection.

It should be pointed out, in conclusion, that the determination of rectifier dissipation is useful in the evaluation of overall efficiency of a connection, and may be used to estimate the voltage drop of a connection, since the total dissipation is approximately equal to the product of the output current and the connection voltage drop.

6. Acknowledgments

The author is indebted to Mr. K. A. Matthews and members of his staff for their advice and assistance in the preparation of this paper.

[12] T. H. Kinman, G. A. Carrick, R. G. Hibberd, and A. J. Blundell, "Germanium and Silicon Power Rectifiers," *Proceedings of the Institution of Electrical Engineers*, part A, volume 103, pages 89–107; April, 1956.

Phototubes Capable of High Current Output*

By SANFORD F. ESSIG

ITT Laboratories; a division of International Telephone and Telegraph Corporation; Fort Wayne, Indiana

THIS paper describes a series of phototubes that have large dynamic range and are capable of high current output. The tubes, used with solid-type scintillators, have become standard for the measurement of high-intensity short-duration gamma radiation. Electrical and optical characteristics are discussed.

• • •

1. Introduction

The high-current phototube has become somewhat of a standard item for measurement of high-intensity gamma radiation of short duration. The development of these tubes has covered a period of more than 9 years; during this period a very-substantial number of tubes have been produced.

The tubes are of coaxial design; that is, the cathode and anode are cylinders with a 1/2.25 dimensional ratio corresponding to a characteristic impedance of 50 ohms. The output, without amplification, is normally applied directly to a 50-ohm coaxial transmission line to a traveling-wave-type oscilloscope for recording.

The tubes are normally used in conjunction with plastic-type scintillators as illustrated in Figure 1. In type *FW-100*, the scintillator is a cylinder surrounding the phototube. In type *FW-104*, light from the scintillator enters the phototube through a circular window. The closed structure in this case is essentially an integrating cavity providing efficient utilization of the light emitted by the scintillator.

2. Description

Figure 2 shows the development of the phototubes over a period of approximately 9 years; this was interspersed with periods of production. Developmental work continues to date and will be described later. The small phototube of biplanar construction in the foreground is an outgrowth of recent developmental work.

The original phototube (furnished by the Naval Research Laboratory), Figure 2A, was largely an improvised tube consisting of a central cathode cylinder mounted on a multipin glass header and surrounded by a wire-gauze anode terminated at the top terminal cap. Vigorous development began in 1951 with a complete redesign aimed at rugged construction. The thinking at the time indicated a completely

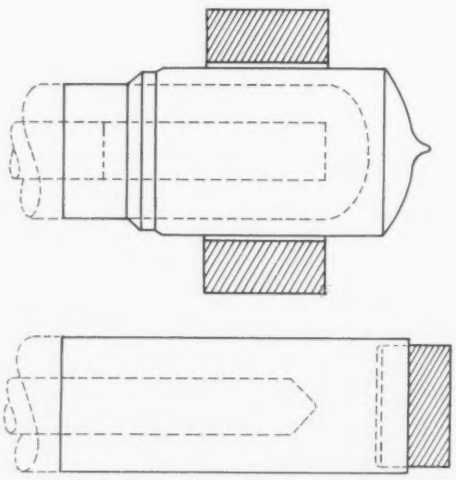


Figure 1—Two types of scintillators (shaded areas).

symmetrical (double-ended) structure with coaxial terminals to permit stacking the tubes as well as to permit attenuation of the electrical wave in the direction opposite to that leading to the recording oscilloscope, the latter to permit the ultimate in electrical response. It is understood that, in practice, neither stacking nor termination has been found advantageous. As a result, types *B* and *C* were discontinued in favor of the half-structure *D*.

The only difference between types *B* and *C* was the all-copper terminal construction of *B*,

* Presented at the Seventh Annual Scintillation Counter Symposium, February 25–26, 1960; Washington, District of Columbia. Reprinted from *IRE Transactions on Nuclear Science*, volume NS-7, numbers 2–3, pages 71–73; June–September, 1960.

The development of these tubes was initiated by ITT Laboratories and the Naval Research Laboratory and since 1952 has been carried on between ITT Laboratories and Edgerton, Germeshausen, and Grier, Incorporated.

which required brazing in a hydrogen atmosphere in contrast with the stainless-steel and Kovar-seam-welded construction that has been standard for years. Tests on type *D* immediately indicated the need for capping the wire-gauze anode to prevent electrical pickup. As a result, type *FW-100* has been the standard for a number of years and only within the past two years has type *FW-104* emerged. Performance indicates that it will probably become the new standard or preferred high-current phototube.

In the biplanar phototubes, of which the *FW-114* is an example, the cathode is a metallic disc mounted on a central post and the anode is an electroformed nickel mesh of 80-percent light transmission in contact with the glass window.

Figure 2—Phototube developmental sequence is shown from left to right.

3. Optical and Electrical Characteristics

Until 1959 the photocathodes were *S4* cesium–antimony type, at which time the cathodes were changed to *S17* for improved sensitivity. Better than 60 percent of the last group of 66 tubes had an actual cathode sensitivity in excess of 40 microamperes per lumen. Attenuation by the wire-gauze anode reduces this by 25 percent. In 1959 work began on application of the multialkali-type cathode to these tubes with an expected sensitivity of 150 microamperes per lumen.

Since all the radiant energy from the scintillator is confined to a relatively narrow spectral band (approximately 350 to 500 millimicrons—the shaded area of Figure 3), cathodes having maximum energy utilization are in the following descending order: (*1*) multialkali on reflective

metal substrate, (*B*) *S17* cesium–antimony on reflective metal substrate, and (*C*) *S4* cesium–antimony on metal substrate. This approximately represents the present direction of effort toward attaining the desired high sensitivity in the violet portion of the spectrum.

If an increase of 66 percent[1] in quantum efficiency at 410 millimicrons can be achieved in the case of antimony–cesium (as data indicates) by using a reflective substrate, then, everything else being equal (which probably cannot be assumed), a similar gain should be attainable by similar treatment of the multialkali cathode. This would result in a quantum efficiency of about 33 percent. Whether this can be achieved still remains unanswered although reasonably good progress has been made to date.

The spectral response curve of the multialkali cathode on reflective substrate, Figure 3, is distinctly different from the translucent *S20* type. It has a characteristic double maximum, the amplitude ratio of the two maximums varying considerably from cathode to cathode. Data taken are insufficient to justify publication of a standard response curve.

In view of the narrow spectral band of the scintillator output, the rating of these tubes in terms of luminous sensitivity will be discontinued in favor of radiant sensitivity at a specified wavelength within the emission band, or over a narrow band in this vicinity.

The energy distribution of a tungsten lamp having a 2870-degree-kelvin color temperature,

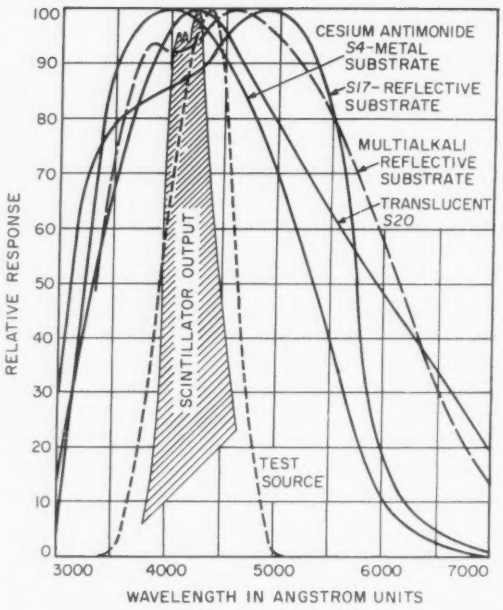


Figure 3—Spectral response of cathodes and their relation to the fluorescence of the scintillator. The tungsten-lamp test source mentioned in the text is a fair approximation to the scintillator response.

filtered by a 2-millimeter-thick Corning *5113* filter, as seen in Figure 3, is reasonably suitable for this measurement.[2]

For maximum linearity at high current, space charge must obviously be avoided by operation at adequate voltage. Inserting the factors $\beta^2 = 0.354$, $r = 4.5$ centimeters, $l = 10$ centimeters, and $V = 10$ kilovolts in the Langmuir space-charge formula for concentric cylinders, a saturation current of 92 amperes is obtained. This is the idealized case with no correction for end effect and disregard of the voltage–current knee, with the result that the expected maximum linear current is considerably less than this value.

On test, where the cathodes

[1] The quantum efficiencies of *S4* and *S11* cathodes are approximately 12 percent at 410 millimicrons; the *S20* and *S17*, approximately 20 percent.

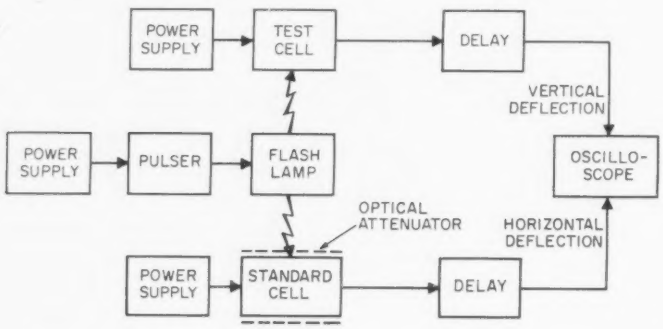


Figure 4—Diagram of Edgerton, Germeshausen, and Grier phototube dynamic test circuit.

[2] One lumen of 2870-degree-kelvin-color-temperature light filtered by such a filter is the equivalent of approximately 137 microwatts. This computed value may later be corrected on better filter calibration.

of *FW-100* tubes were illuminated from one side only, the maximum measured current before departure from linearity has been 55.5 amperes. (The average measured value is approximately 45 amperes.) Under conditions of complete cathode illumination, this value should be considerably higher. Theory and observation therefore check reasonably well.

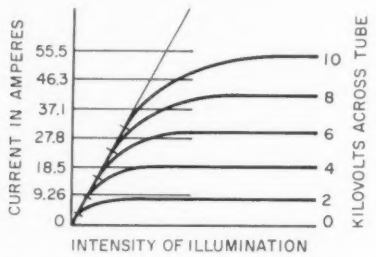


Figure 5—Tracing of a typical family of curves for a tube. The vertical scale is current in amperes and each curve is marked with the voltage applied to the tube.

It is seen that the useful dynamic range of the tube is fixed largely by the external recording circuits rather than by the tube itself.

Figure 4 is a block diagram of the Edgerton, Germeshausen, and Grier dynamic test circuit, which essentially compares the output of the tube on test with that of an optically attenuated standard tube, the latter assuring linearity of output with respect to light input. A pulse of light falls simultaneously on both tubes, the outputs being applied respectively to the vertical and horizontal deflection system of a recording oscilloscope. Figure 5 is a typical family of traces at a series of indicated operating voltages.

Figure 6 shows the characteristic spread in dark current versus voltage for a large number of *FW-100* tubes. The source of the dark current is principally field emission. While the specified maximum dark current is 10^{-8} ampere at 8 kilovolts, it will not exceed 10^{-9} ampere at 10 kilovolts in approximately half of the more-recent tubes.

Figure 6 also shows the characteristic current-versus-voltage plateau at a nominal current of approximately 10^{-7} ampere. If the cathode is illuminated with violet light, the plateau is almost perfectly flat over a wide voltage range, while with white light the slope is considerable. With light of only the longer wavelengths, the effect is still more pronounced. This is attributed to the porous cathode or a lowering of the potential barrier by the electric field. Due to a preponderance of low-velocity electrons emitted at the longer wavelengths, the cathode contributes more electrons as the field reaches deeper into the cathode lattice. Conversely, with a greater population of high-velocity electrons, current is essentially independent of voltage over a wide range before the effect is observed.

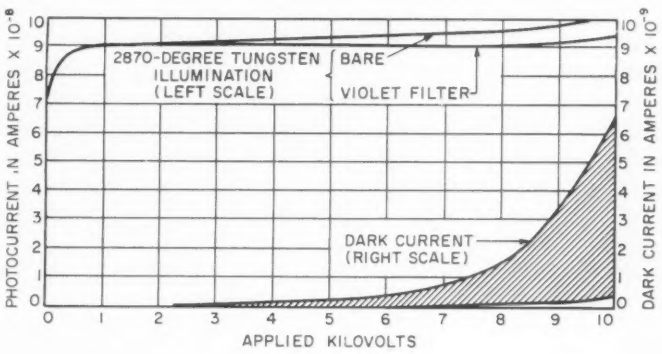


Figure 6—Spread from tube to tube of dark current versus voltage.

While extensive data of the small biplanar *FW-114* cell (Figure 2) have not been taken to date, preliminary tests indicate that it is linear up to 25 amperes at 10 kilovolts. It, too, is of rugged construction and promises to be a useful addition to the series.

4. Acknowledgments

Information and data furnished by Mr. J. V. Shipman of the Naval Research Laboratory and Messrs. R. B. Patton and P. A. Hawkins of Edgerton, Germeshausen, and Grier, Incorporated, is respectfully acknowledged.

AN/MRC-55 General-Purpose Mobile High-Frequency Communication Equipment

By M. C. DETTMAN

ITT Federal Division; Clifton, New Jersey

MILITARY communication set *AN/MRC-55* is a general-purpose high-frequency equipment designed for amplitude-modulated voice, continuous wave, frequency-shift-keyed teleprinter, or simultaneous voice and teleprinter transmission and reception. It is jeep mounted and consists of a nominal 100-watt 2-to-30-megacycle-per-second transmitter, two identical receivers with 2-to-32 megacycle-per-second tuning range, suitable control boxes for local, remote, and relay operation, and all accessories such as microphone, headset, handset, telegraph keys, writing table, loudspeaker, and antennas. The need for extensive operator training has been eliminated by utilizing techniques that automatically perform transmitter-tuning and antenna-loading functions. The radio set is designed for deep-water fording and may be operated from either the 28-volt direct-current supply of the vehicle or from a 115-volt 60-cycle-per-second single-phase source.

Figure 1—At top, *AN/MRC-55* high-frequency radiotelephone and radiotelegraph station mounted in jeep.

As performance specifications for electronic equipment become more severe, generally resulting in more-complex apparatus, the problems of operator training and field maintenance are also multiplied. The *AN/MRC-55* mobile communications set, shown mounted in its carrier vehicle in Figure 1, is designed to minimize these problems and at the same time provide superior communication in the high-frequency range over that obtainable with prior equipments. Operator training time has been reduced to a negligible amount. During evaluation testing, it was determined that untrained personnel could be taught proper operation of the complete equipment in 30 minutes. The automatic tuning features incorporated in the transmitter have made this possible. The techniques used ensure that optimum transmitter power is coupled to the radiating antenna, resulting in superior range. We need no longer be plagued by the problem of improper manual tuning for a maximum indication on a radio-frequency ammeter, which may only mean maximum circulating current in the antenna tuner and minimum radiated power. Modular plug-in chassis construction is employed in both receiver and transmitter to expedite field maintenance. Numerous test points and front-panel monitoring of important transmitter power and signal circuits are provided.

Figure 2—Receiver *R808/GRC-14*.

The nomenclature, *AN/MRC-55*, is applied to the complete radio station, including the vehicle, an *M-38A1* modified for deep-water fording along with a special high-output, battery–generator system, *MK-366/MRC-55*. The nomenclature *AN/GRC-14* applies to the radio equipment only as employed for general ground use (fixed installation).

1. Radio Receiver

The radio receiver, *R-808/GRC-14*, is shown in Figure 2. It is a high-performance superheterodyne receiver covering a frequency range of 2 to 32 megacycles per second in four bands. Double conversion is used above 8 megacycles per second. It is capable of receiving continuous waves, telephone, and frequency-shift-keyed signals, or the latter two simultaneously. The converter for frequency-shift keying is integral with the receiver, so that the teleprinter loop circuit can be keyed directly. It uses either a high-impedance antenna or a 50-ohm coaxial

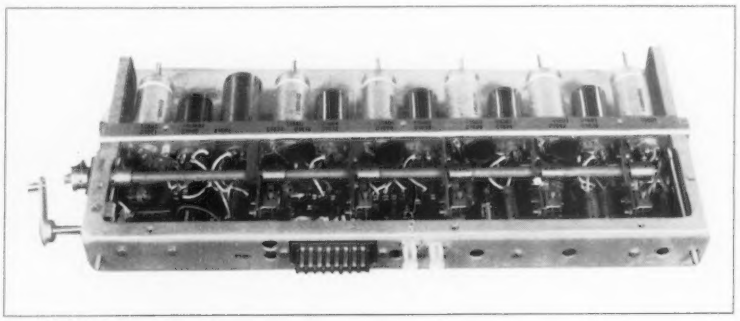


Figure 3—Intermediate-frequency amplifier.

input. Sensitivity is better than 3 microvolts for a 20-decibel signal-plus-noise-to-noise ratio for continuous-wave signals and 5 microvolts for a 10-decibel signal-plus-noise-to-noise ratio with a 30-percent-modulated carrier over the 2-to-8-megacycle-per-second range. Above 8 megacycles per second, the sensitivity becomes 5 microvolts for the continuous-wave case and 8 microvolts for modulated-continuous-wave operation. Two degrees of intermediate-frequency bandwidth

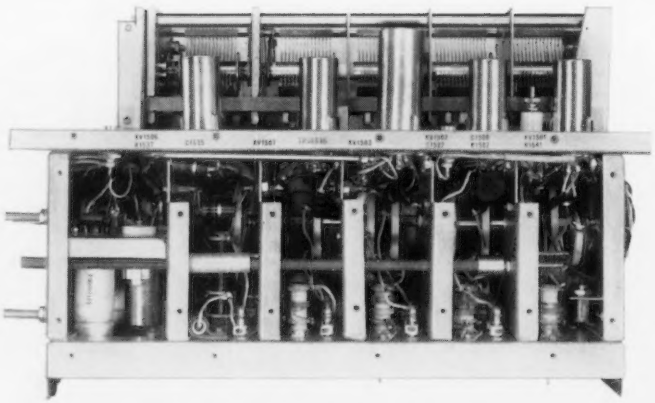


Figure 4—Tuning unit of receiver.

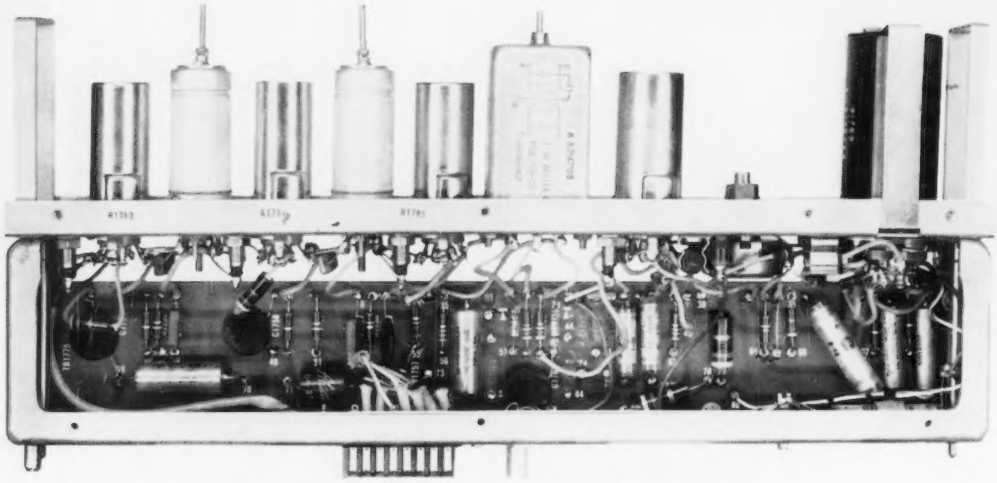


Figure 5—Receiver module for frequency-shift-keying converter.

are provided, namely 3.5 and 7.0 kilocycles per second at 6 decibels down. The 6-to-60-decibel bandwidth ratio is better than 3.5.

The intermediate-frequency amplifier is shown in Figure 3. A combination of printed-circuit and point-to-point wiring is used. Two degrees of audio selectivity are also provided by including a band-pass filter with a 350-cycle-per-second passband centered at 1000 cycles per second. The use of two radio-frequency preselector stages, double conversion above 8 megacycles per second and careful attention to circuit wiring, layout, and filtering have resulted in an intermediate-frequency rejection of better than 60 decibels, an image rejection

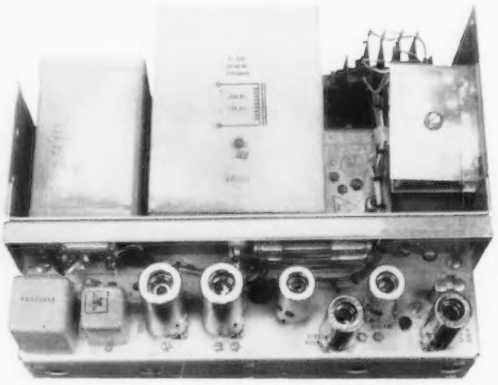


Figure 6—Receiver power supply and amplifier module.

of better than 55 decibels at 2 megacycles per second, reducing to 25 decibels at 32 megacycles per second, and minimum local-oscillator radiation. Figure 4 shows an interior side view of the radio-frequency head. The main tuning scale is approximately 105 inches (267 centimeters) long, allowing incremental scale markings of 2, 5, 10, and 20 kilocycles per second for the four bands from lowest to highest. The scale can be calibrated against a built-in crystal calibrator at check points every 200 kilocycles per second. A high degree of frequency and gain stability is achieved through the use of temperature-compensated components in the high-frequency oscillator, as well as utilization of a transistor-type voltage regulator that supplies all plate and screen voltages and all tube heaters except the frequency-shift-keying converter-unit heaters. Miniature tubes are used in all circuits except the audio output stage, frequency-shift-keying polar-relay drive, and the power supply, which use transistors. Figures 5 and 6 show the frequency-shift-keying converter module and the amplifier and power-supply module, respectively. Each of the two identical receivers uses 23 tubes, 8 transistors, and 6 semiconductor diodes. The receivers are of sealed and gasketed construction and may be submerged during deep-water fording operations.

2. *Radio Transmitter*

Radio transmitter *T-631/GRC-14* is shown in Figure 7. It is rated at a nominal 100 watts of carrier power output over the frequency range of 2 to 30 megacycles per second. It will provide continuous-wave, radiotelephone, frequency-shift-keyed, or simultaneous telephone and frequency-shift-keyed signals. The output circuit normally energizes the jeep whip antenna but may optionally be connected to a 50-ohm coaxial line or to a long wire antenna. Tuning of the radio-frequency amplifier circuits and antenna loading are accomplished automatically. Frequency accuracy and stability are determined by a digitally set synthesizer, which is used to control the frequency of an adjustable-frequency oscillator. The variable-frequency oscillator is manually adjusted to the desired frequency by means of a synchronization indicator that shows when the automatic frequency control is acting on the oscillator frequency. Frequency increments in 100-cycle-per-second steps are available in the stabilized condition. In the event of failure of the synthesizer circuits, the variable-

Figure 7—Transmitter *T-631/GRC-14*.

frequency oscillator operates as a standard master oscillator. The output of the variable-frequency oscillator drives the automatically tuned buffer and power amplifier circuits. These, in turn, are connected to the automatically tuned antenna-matching circuit. The transmitter consists of the main chassis and 8 plug-in units. A top view of the transmitter is shown in Figure 8 and the bottom view in Figure 9. All the plug-in units are readily replaceable for rapid field servicing when required. A total of 29 tubes and 21 semiconductor diodes are utilized in the transmitter.

The power output stage utilizes a *4X250F* tetrode. This stage is high-level modulated by two *4X250F* tetrode tubes for telephone operation. An interesting feature of the radio-frequency amplifier–modulator units is the employment of motor-driven adjustable inductances utilizing fine silver ribbon wire for the conductor. The active portion of the inductance is on a low-loss grooved dielectric form, while the inactive portion is wound on a gold-plated short-circuiting drum. Band changing is accomplished by switching fixed capacitors. The use of moulded glass-dielectric capacitors and precision ceramic capacitors resulted in a reduction in size and weight. Space saving was accomplished in the radio-frequency tuner by using a similar adjustable inductance and a vacuum-dielectric adjustable capacitor. These features are clearly shown in Figure 8.

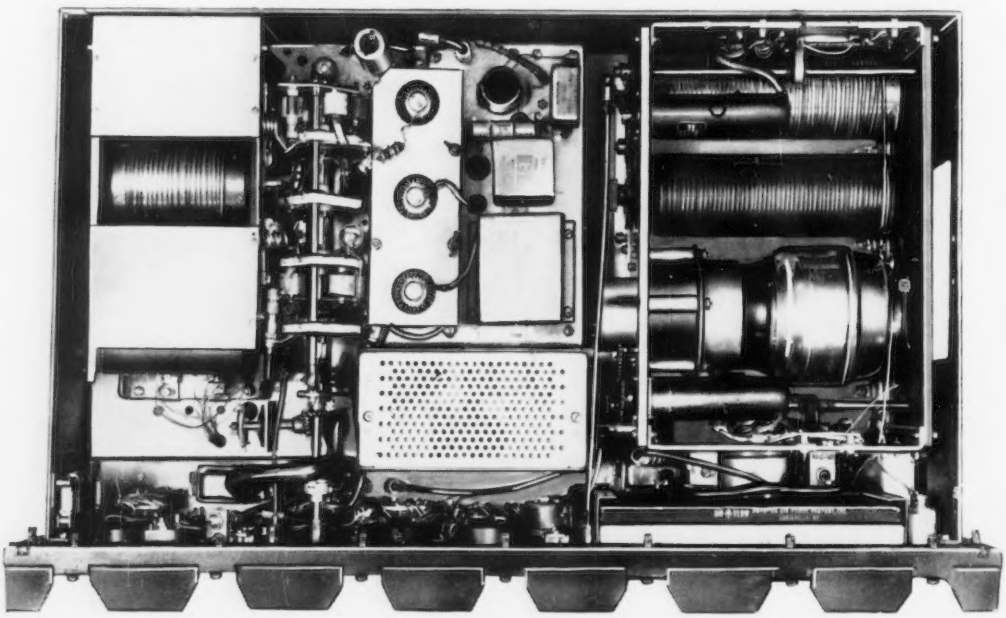


Figure 8—Top deck of transmitter. The radio-frequency tuner is at the right, the audio-frequency amplifier is beneath the perforated plate, and just back of it the anode terminals of the amplifier and modulator tubes may be seen through circular holes in the clamping plate. The circuits for the amplifier tube are at the left.

2.1 Frequency Synthesis

A block diagram of the frequency synthesizer is shown in Figure 10. The variable-frequency oscillator establishes the frequency of transmission. The technique employed to stabilize the transmitted frequency consists of mixing the output frequency from the variable-frequency oscillator with various crystal-controlled frequencies in such a manner that a frequency between 2 and 3 megacycles per second is obtained. This frequency is then compared in a phase detector with a signal of the same frequency obtained by mixing the output of another crystal-controlled oscillator and the output of a high-stability interpolation oscillator. The phase-

detector output is a direct voltage that controls the effective capacitive reactance of back-biased diodes that are used to reactance modulate the variable-frequency oscillator and thus control its frequency. Frequency-control action may be best illustrated by a specific example.

In the following discussion, all references are to Figure 10. If we choose an operating frequency of 23.645 300 megacycles per second, the following actions take place. The frequency selector dials are set to read 23.645 300 megacycles per second and the variable-frequency oscillator is adjusted to show synchronization.

The decade switch in Figure 10 is controlled by the tens-of-megacycles-per-second digit of the frequency selector dials. When set at 2, corresponding to 20's of megacycles per second, in addition to making the connections shown, it de-energizes the mixer-selector relay and selects the 20-to-30-megacycle-per-second band of the variable-frequency oscillator.

When the units-of-megacycles-per-second selector dial is set at 3, through paths not shown in the figure, it sets the variable-frequency oscillator to the 23-megacycle-per-second range, selects a 14-megacycle-per-second signal from the fundamental oscillator, and a 35-megacycle-per-second signal from the oscillator–tripler, which two frequencies are applied to mixers *A* and *B*, respectively. The oscillator-doubler is inoperative for these settings.

The variable-frequency-oscillator frequency of 23.645 300 megacycles and the 14-megacycle wave from the fundamental oscillator produce in mixer *A* the sum frequency of 37.645 300 megacycles. This passes through the band-pass filter and via the contacts of the mixer-selector relay to mixer *B*. The 35-megacycle frequency from the oscillator–tripler passes through the decade switch to mixer *B*. These two frequencies produce in mixer *B* a difference frequency of 2.645-300 megacycles per second, which goes through the band-pass filter to the phase detector. In a similar manner, all other operating frequencies are converted to the 2-to-3-megacycle range, except for frequencies in the 2-to-3-megacycle band that pass directly through mixer *B* to the band-pass filter and the phase detector.

We now proceed to another signal chain that develops the comparison signal from which the

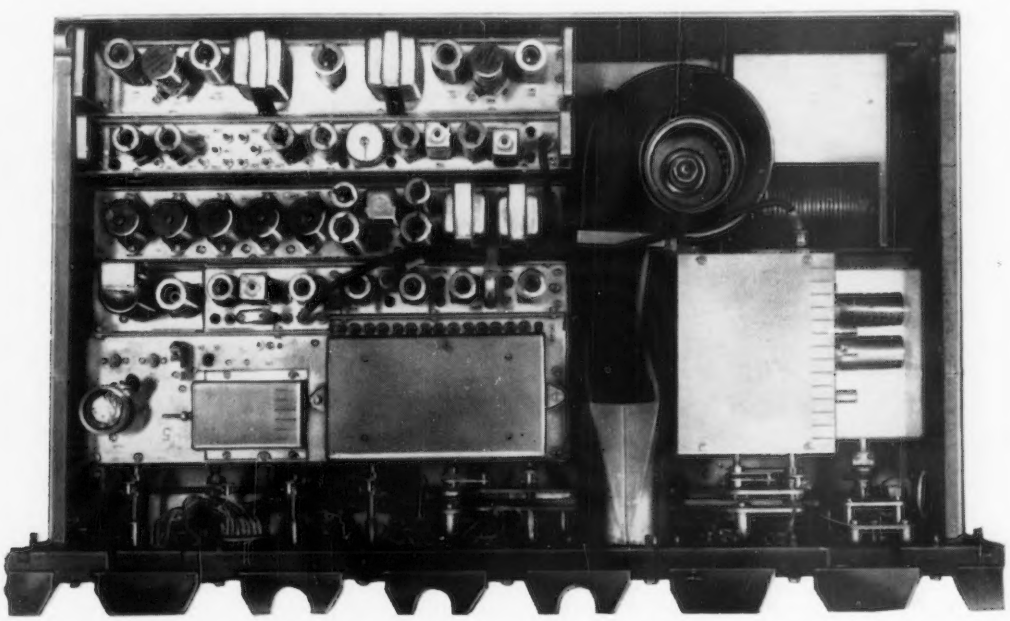


Figure 9—Bottom deck of transmitter. The variable-frequency oscillator is in the box to the right of the blower. The four chassis at the left from the front to the back are the reference-frequency oscillators and mixer keying circuits, mixer-stabilizer, and servocontrol circuits.

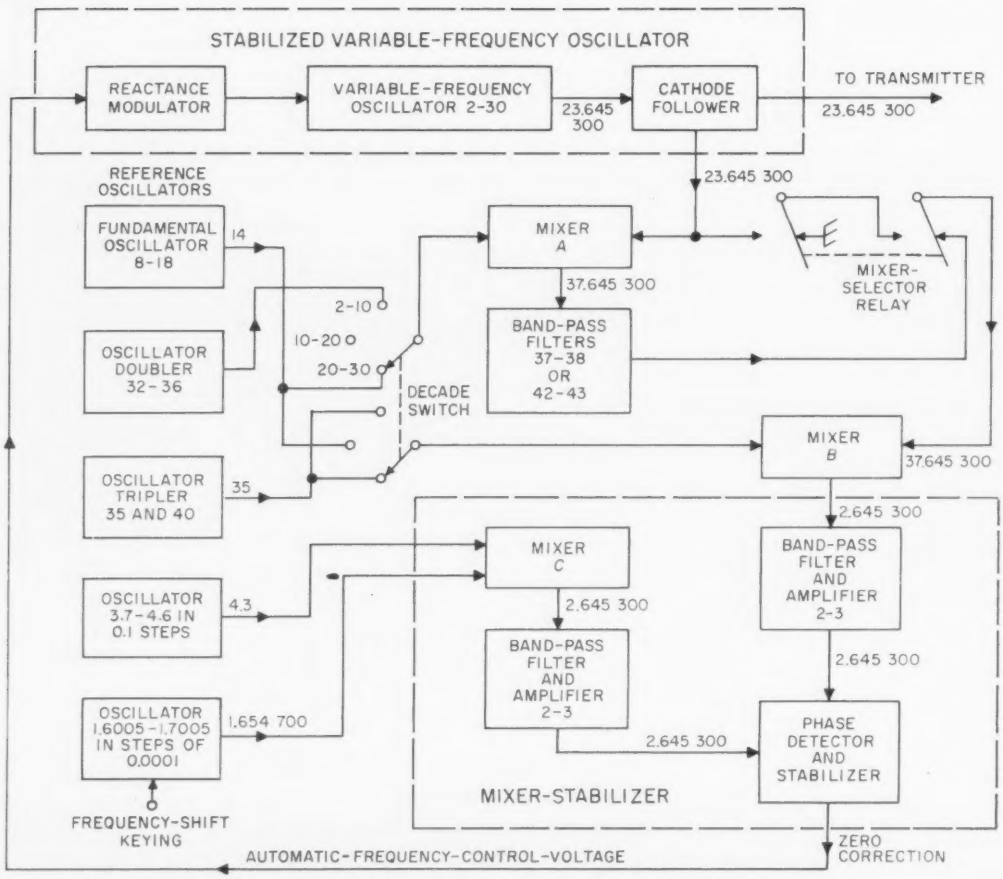


Figure 10—Frequency synthesizer for the transmitter. The numbers are the frequency ranges in megacycles per second.

Figure 11—The handwheel-clamped access doors of the power supply are closed for submerging in water and must be opened for ventilation when operating.

phase detector produces any necessary error signal to correct the variable-frequency oscillator.

When the hundreds-of-kilocycles decade dial of the frequency selector is set to 6, a 4.3-mega-

oscillator to make the two frequencies impressed on the phase detector the same.

The precision of the output frequency depends on the accuracy of the crystal and 1.6005-to-

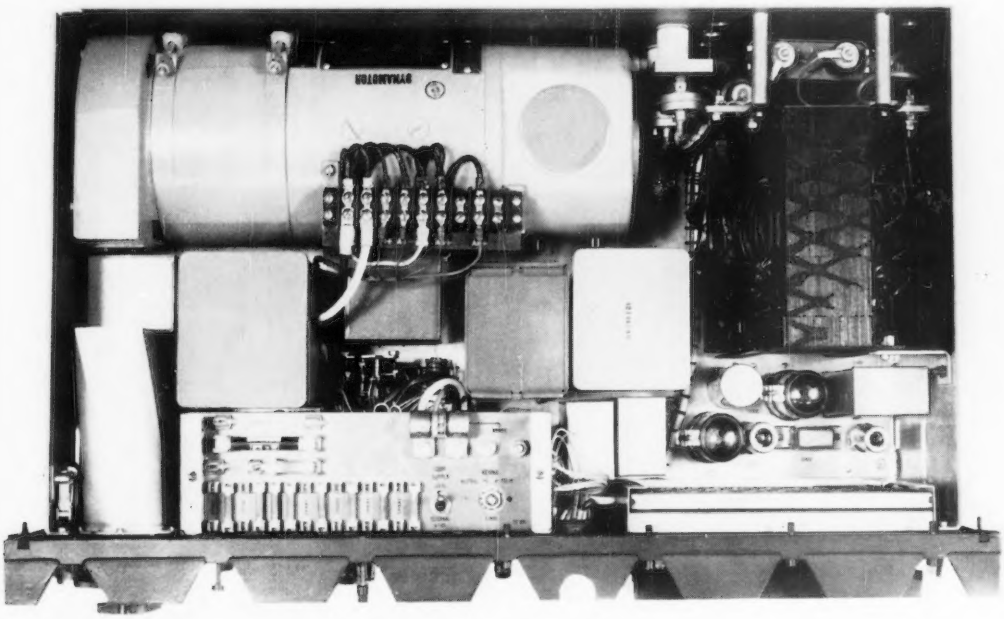


Figure 12 —The transformer and silicon rectifiers in the upper right corner permit the dynamotor of this power supply to be operated from alternating current.

cycle-per-second signal is selected from the oscillator that produces 100-kilocycle-per-second steps between 3.7 and 4.6 megacycles per second. The next three decades of the frequency selector set the frequency of the 1.6005-to-1.7005-megacycle-per-second oscillator, which is adjustable in 100-cycle-per-second steps over its range, to 1.654 700 megacycles per second. In mixer *C*, these two frequencies produce a difference frequency of 2.645 300 megacycles per second, which goes through the band-pass filter to the phase detector.

The phase detector compares the two frequencies that reach it. If they are the same, no output voltage is produced to change the variable-frequency-oscillator. If the two frequencies are not the same, an error voltage is produced by the phase detector of such polarity and amplitude as will act on the reactance modulator to adjust the frequency of the variable-frequency

1.7005-megacycle-per-second reference oscillators. In the event of failure of any of the synthesizer units, the transmitter will continue to operate under control of the variable-frequency oscillator but without benefit of frequency stabilization.

2.2 Power Supply

The power supply designated *PP-1711/GRC-14* is shown in Figures 11 and 12, furnishes all necessary voltages for transmitter operation, as well as power for the teleprinter and the associated loop circuit. The unit is designed to operate from a primary power source of either 28 volts at 40 amperes of direct current or from a 110-volt single-phase 60-cycle-per-second alternating-current supply at 1.6 kilovolt–amperes.

The main power switch selects the type of power source. In direct-current operation, the dynamotor is supplied directly from the primary

power source. For alternating-current operation, primary power goes to the step-down transformer shown in the upper right-hand corner of Figure 12 and then to the two silicon rectifiers immediately adjacent. These rectifiers are connected for full-wave operation and are mounted on a heavy heat sink cooled by the input air stream to the dynamotor. The unfiltered direct current supplies the dynamotor input. The output voltages of the dynamotor are filtered. The voltage-sensitive circuits are operated from an intermediate value of voltage for which regulation is provided.

The receiver power supplies are self-contained. They also operate from either 28 volts direct current or 110-volt, 60-cycle-per-second, single-phase alternating current, drawing about 80 watts in each case.

2.3 Accessories

The control group, *O.1-1444/GRC*, is shown in Figure 13 and consists of control box *C-2171/GRC-14* installed at the transmitter and control box *C-2172/GRC-14* located in the field. These units provide for remote operation over a mile (1.6 kilometers) of two-wire line and also supply standard two-way telephone and ringing facilities so that the local operator at the jeep can talk with the remote operator. The transmitter can be turned on and off and modulated with microphone or keying signals from the remote point.

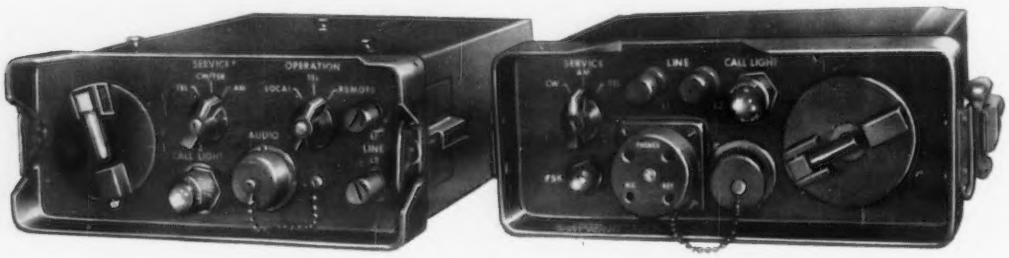


Figure 13—At left is control unit *C-2171/GRC-14* installed at the transmitter and at the right is *C-2172/GRC-14*, which may be at the end of a mile (1.6 kilometers) of line, for remote operation.

Included in the accessories are all necessary microphones, handsets, headsets, and special wrenches. An operator's writing table is supplied.

In addition to the two whip antennas, normally installed on the jeep, a long wire antenna is also provided.

United States Patents Issued to International Telephone and Telegraph System; November 1959—January 1960

BETWEEN November 1, 1959 and January 31, 1960, the United States Patent Office issued 104 patents to the International System. The names of the inventors, company affiliations, subjects, and patent numbers are listed below.

R. T. Adams and J. B. Harvey, ITT Laboratories, Pulse Detector Responsive to Both Pulse Amplitude and Duty Cycle, 2 916 618.

H. H. Adelaar and F. C. L. de Wit, Bell Telephone Manufacturing Company (Antwerp), Electrical Identification System, 2 913 531.

D. F. Allison, ITT Laboratories, Electroforming of Semiconductive Cells, 2 915 687.

J. Augustin and O. Holstein, C. Lorenz (Stuttgart), Receiving System for Telegraph Signals, 2 911 471.

M. C. E. Bataille and G. M. Planchais, Laboratoire Central de Télécommunications (Paris), Electronic Finder, 2 919 309.

J. K. Bates, Jr., ITT Laboratories, Pulse Width and Repetition-Rate Discriminator, 2 912 579.

A. H. W. Beck and G. P. de Mengel, Standard Telephones and Cables (London), Electron Velocity-Modulation Tubes, 2 918 599.

A. H. W. Beck, A. B. Cutting, A. D. King, and A. D. Brisbane, Standard Telephones and Cables (London), Thermionic Cathodes, 2 912 611.

J. I. Bellamy, Kellogg Switchboard and Supply Company, Fast-Acting Electromagnetic Counting Device, 2 909 712.

W. Berthold, C. Lorenz (Stuttgart), Gun System Comprising an Ion Trap, 2 921 212.

W. Berthold, C. Lorenz (Stuttgart), Signal-Storage Tube, 2 918 601.

N. A. Blake, ITT Industrial Products Division, Blanking System, 2 911 469.

M. C. Branch, Standard Telephone and Cables (London), Pulse Transmitting Circuit, 2 914 667.

F. H. Bray, P. M. King, and G. C. Hartley, Standard Telephones and Cables (London), Line Identification in Telecommunication Systems, 2 913 530.

F. H. Bray and R. G. Knight, Standard Telephones and Cables (London), Electric Signal Repeaters, 2 914 615.

F. H. Bray and R. G. Knight, Standard Telephones and Cables (London), Electrical Signaling Systems, 2 911 475.

W. T. Brown, Farnsworth Electronics Company, High-Speed Waveguide Switch, 2 917 719.

J. H. Brundage, ITT Laboratories, Signal-Transmission Systems, 2 920 291.

H. N. Capen, C. R. Brown, and L. F. Evans, ITT Laboratories, Method of Manufacturing a Coaxial Filter, 2 911 333.

D. L. Carpenter and L. E. Clements, Farnsworth Electronics Company, Automatic Automotive Headlight Dimming Circuit, 2 917 665.

W. F. S. Chittleburgh, T. H. Walker, and A. Jessop, Standard Telecommunication Laboratories (London), Telegraph Regenerative Repeaters, 2 918 529.

J. E. Cox and G. L. Hasser, Kellogg Switchboard and Supply Company and Federal Telephone and Radio Company, Revertive Call-Selector Circuit, 2 917 585.

J. Dascotte, Le Matériel Téléphonique, Transmission System for Radioelectric Waves, 2 921 274.

D. W. Davis, ITT Laboratories, Storage Cathode-Ray Tube, 2 916 661.

E. C. L. deFaymoreau, ITT Laboratories, Navigation System with Beacon Identification, 2 916 737.

E. C. L. deFaymoreau, ITT Laboratories, Potentiometer, 2 912 662.

E. C. L. deFaymoreau, ITT Laboratories, Radionavigation System, 2 912 689.

E. C. L. deFaymoreau and M. Mandel, ITT Laboratories, Searching and Tracking Receiver, 2 916 738.

E. C. L. deFaymoreau, F. Singer, Jr., and H. R. Wolter, ITT Laboratories, Test Set, 2 912 688.

M. den Hertog and C. deZeeuw, Bell Telephone Manufacturing Company (Antwerp), Scanning Circuit, 2 913 528.

M. den Hertog and C. deZeeuw, Bell Telephone Manufacturing Company (Antwerp), Telecommunication Exchange System, 2 912 510.

A. R. Denz, Kellogg Switchboard and Supply Company, Temperature-Compensated Direct-Current Amplifier, 2 918 627.

L. A. deRosa, M. Masonson, and F. J. Lundberg, ITT Laboratories, Radionavigation Beacon, 2 915 750.

R. M. Dobrowolski and D. C. Seeley, ITT Laboratories, Selenium Rectifier, 2 912 353.

S. H. M. Dodington, ITT Laboratories, Pulse Transmitting and Receiving System using a Common Source of Oscillations, 2 916 614.

S. H. M. Dodington and B. B. Mahler, ITT Laboratories, Distance-Measuring Equipment, 2 912 686.

J. L. J. Donceel and H. C. J. Van Der Harck, Bell Telephone Manufacturing Company (Antwerp), Electromagnetic Transducer, 2 914 621.

P. E. Dorney and D. M. Sharp, ITT Laboratories, Microwave Sweep Generators, 2 912 652.

E. H. Eberhardt, ITT Laboratories, Ion Gauge, 2 913 630.

P. T. Farnsworth, Capehart–Farnsworth Company, Color Television Apparatus, 2 921 228.

L. G. Fischer, ITT Laboratories, Sector-Scanning Direction-Finder Systems, 2 912 692.

A. R. Gobat, ITT Laboratories, Method of Cutting Semiconductive Material, 2 911 773.

F. P. Gohorel, A. J. Henquet, and J. J. Perrot, Compagnie Générale de Constructions Téléphoniques and Le Matériel Téléphonique (Paris), Marker Control for Crossbar Automatic Telephone System, 2 911 477.

D. D. Grieg and H. F. Engelmann, ITT Laboratories, Microwave Transmission Lines, 2 915 718.

G. L. Hall, ITT Laboratories, Electron-Beam Gun Systems, 2 916 666.

H. Havstad, ITT Laboratories, Frequency Multiplier, 2 920 286.

J. A. Henderson and N. E. Hoag, ITT Laboratories, Photoemissive Tube, 2 913 610.

A. J. Henquet, Le Matériel Téléphonique (Paris), Potential Comparing Systems, 2 921 234.

A. J. Henquet and J. J. Perrot, Le Matériel Téléphonique and Compagnie Générale de Constructions Téléphoniques (Paris), Switching System Particularly Applicable to Automatic Telephone System, 2 913 534.

J. Hessler, Farnsworth Electronics Company, Twin-Probe Waveguide Transition, 2 909 735.

R. B. Hoffman and T. E. Ahlstedt, Federal Telephone and Radio Company, Mulitplex Transmission System, 2 918 532.

R. W. Hughes, ITT Laboratories, Communication System, 2 912 506.

R. W. Hughes, ITT Laboratories, Crosstalk Suppression, 2 912 509.

R. W. Hughes, ITT Laboratories, Pulse Communication System, 2 912 507.

R. W. Hughes, ITT Laboratories, Repeater Station for a Pulse Multiplex System, 2 912 508.

H. L. Hull, Farnsworth Electronics Company, Method of Fabricating a Phosphor Screen, 2 920 959.

D. S. Humphreys, Farnsworth Electronics Company, Audio Amplifier System, 2 922 846.

R. W. Hunter, Farnsworth Electronics Company, Electron Discharge Devices, 2 917 658.

R. W. Hutton, A. T. Sigo, and C. J. Adams, Kellogg Switchboard and Supply Company, Multigroup Direct-Access Crossbar Telephone Switching System, 2 909 611.

C. E. Jones, Jr., ITT Laboratories, Data Processing System, 2 916 727.

R. V. Judy, Kellogg Switchboard and Supply Company, Multiexchange Telephone System, 2 909 612.

W. Klein and W. Frix, C. Lorenz (Stuttgart), Attenuation for Traveling-Wave Tube, 2 911 599.

W. Klein and W. Fritz, C. Lorenz (Stuttgart), Traveling-Wave Tube, 2 921 225.

J. J. B. Lair, ITT Laboratories, Pulse Modulation Systems, 2 920 288.

H. S. Leman, Standard Telephones and Cables (London), Telephone Signaling Systems, 2 916 558.

J. K. Lennard, ITT Laboratories, Electron Discharge Device, 2 913 609.

J. A. Leno, Standard Telecommunication Laboratories (London), Electric Semiconductor Devices, 2 913 643.

G. X. Lens, Bell Telephone Manufacturing Company (Antwerp), Combined Code Recorder and Selector, 2 914 197.

E. J. Leonard, A. Hemel, H. Tarschisch, K. A. Karow, J. C. Gibson, and K. L. Liston, Kellogg Switchboard and Supply Company, Register-Translator Crossbar Telephone System, 2 918 533.

H. L. Levin, ITT Laboratories, Electron-Gun Structure, 2 916 649.

A. Lieb, C. Lorenz (Stuttgart), Fluorescent Screen for Electron-Ray Tubes Operating at Low Voltage, 2 921 201.

A. Linnebach, C. Lorenz (Stuttgart), Ring Filter Circuit, 2 919 417.

C. Lucanera and H. Avery, ITT Laboratories, Omnidirectional Beacon Antenna, 2 912 693.

F. J. Lundberg, ITT Laboratories, Radio-Frequency Delay Line, 2 916 615.

F. T. Machalek, Farnsworth Electronic Company, Automatic Headlight System, 2 921 232.

F. T. Machalek, Farnsworth Electronics Company, Headlight Overriding System, 2 912 621.

M. Mandel, ITT Laboratories, Aerial-Navigation System, 2 912 690.

M. Mandel, ITT Laboratories, Radionavigation Receiver, 2 912 691.

T. J. Marchese and R. W. Wilmarth, ITT Laboratories, Pulsed-Cathode Electron Gun, 2 912 616.

C. Marteleur, Mix & Genest (Stuttgart), Magnetic Recording and Reproducing Device, 2 916 559.

F. P. Mason and J. L. Stevens, Creed & Company (Croydon), Power-Supply Equipment, 2 916 682.

S. E. Mayer, Standard Telephones and Cables (London), Low-Resistance Contacts to Germanium, 2 914 449.

O. S. Meixell, ITT Laboratories, Amplifier System, 2 920 230.

M. Muller, C. Lorenz (Stuttgart), Magnetic Fluxmeter, 2 918 622.

B. J. Norris and F. R. Noll, Farnsworth Electronics Company, Transistor Voltage Regulators, 2 922 945.

H. Oden, V. Epinatjeff, K. Sass, and H. Hehr, Mix & Genest (Stuttgart), Pulse Transmitter System, 2 919 310.

R. K. Orthuber and C. G. Larson, Farnsworth Electronics Company, Image Device, 2 909 667.

R. L. Plouffe, Jr., ITT Laboratories, Pulse Communication System, 2 920 142.

J. Polyzou and J. L. Jatlow, Federal Telephone and Radio Company, 2 913 711.

D. C. Rogers, Standard Telephones and Cables (London), Traveling-Wave Tubes, 2 918 593.

N. Rouche, Bell Telephone Manufacturing Company (Antwerp), Calculating Means, 2 911 149.

M. Scata and B. Dal Bianco, Fabbrica Apparecchiature per Communicazioni Elettriche Standard (Milan), Sealed Oscillating Member or Armature of Polarized Electromagnetic Relay or Switch, 2 921 161.

H. Schroder, C. Lorenz (Stuttgart), Polarized Relay, 2 911 574.

A. T. Sigo, Kellogg Switchboard and Supply Company, Flat-Bank Crossbar Switch, 2 909 613.

M. T. Smid and A. Becker, ITT Laboratories, Quick-Acting Fastener Assembly, 2 910 756.

R. G. Stemp and M. H. Lloyd, Creed & Company (Croydon), Facsimile Apparatus, 2 914 372.

H. F. Sterling, Standard Telecommunication Laboratories (London), Refining Processes for Semiconductor Materials, 2 914 397.

I. R. Studebaker, Farnsworth Electronics Company, Cathode-Ray-Tube System, 2 913 623.

J. F. Sullivan and G. W. Reich, Jr., ITT Laboratories, Synchronized Data-Processing System, 2 912 585.

D. L. Thomas, Standard Telephones and Cables (London), Protection Against Failure of Pilot Wave in Carrier Communication System, 2 921 267.

S. R. Wagner, ITT Laboratories, Fluid-Cooled Electronic Chassis, 2 912 624.

W. K. Weston, C. W. Clarke, and R. J. M. Andrews, Standard Telephones and Cables (London), Submarine-Cable Repeater Housings, 2 914 599.

E. P. G. Wright, Standard Telecommunication Laboratories (London), Electric Telegraph Systems, 2 918 526.

E. P. G. Wright and J. Rice, Standard Telecommunication Laboratories (London), Telecommunication Exchange Systems, 2 913 527.

Data Processing System

2 916 727

C. E. Jones, Jr.

A data processing system is described that provides a cathode-ray-tube display of the data being transmitted. The data to be processed are repeatedly scanned at high speed by a flying-spot scanner. The scanned data are applied to a cathode-ray-tube indicating monitor and repeated at a sufficiently high rate to maintain the display. At each scan, a portion of the data is transmitted to a slower data processing equipment.

Low-Resistance Contacts to Germanium

2 914 449

S. E. Mayer

A method is disclosed for making low-resistance contacts to a germanium crystal by depositing metastable nickel stannide on the germanium surface by electroplating. The deposit is heated to a temperature of 250 to 300 degrees centigrade to release tin from the stannide and this tin surface is then coated with a soft solder to which connections can be made.

Thermionic Cathodes

2 912 611

A. H. W. Beck, A. B. Cutting, A. D. King, and A. D. Brisbane

This invention consists of an indirectly heated dispenser-type cathode made of a sintered mass of mixed powders of metallic nickel, thermionically emissive alkaline-earth carbonates, and a reducing agent, preferably zirconium. This mass is supported on an indirectly heated structure to which it is firmly united. The mixed powders contain from 10 to 40 percent of the carbonate and a proportion not exceeding 1.5 percent of the reducing agent, the remainder being pure nickel.

Method of Cutting Semiconductive Material

2 911 773

A. R. Gobat

To avoid waste of semiconductor material in cutting dice from a block of germanium or silicon crystal, a series of crossed shallow intersecting cuts are made in one surface of the block. This avoids the difficulties of sawing off a complete slab, which requires a relatively wide saw cut, and then separately supporting this slab while cutting it into smaller units.

Transmission System for Radioelectric Waves

2 921 274

J. Dascotte

A simple structure has been invented for rotating the fields of a wave when coupling two rectangular waveguides having their longer dimensions at right angles. The coupling unit is in the form of an isosceles right triangle having the two sides forming the right angle in contact with and of the same dimensions as the longer axes of the two waveguides.

Magnetic Recording and Reproducing Device

2 916 559

C. Marteleur

A magnetic recording device is described in which recorded information can be repeatedly read without demagnitization. This recorder includes a core of magnetic material having two pole faces. A recording is made by applying a magnetizing field to a coil around this core. An element of magnetic material is positioned between the two pole pieces. By vibrating this element, a current can be induced into the magnetizing coil to read the stored information.

In Memoriam

SIR FRANCIS BRAKE

FRANCIS J. E. BRAKE was born in Hong Kong on December 10, 1889. He attended Mountjoy School in Dublin and Saint Pauls School in London.

He trained as an apprentice in the National Telephone Company from 1907 to 1911. He joined the Western Electric Company the following year and rose to chief equipment engineer by 1919. He became managing director of its associate company in Norway in 1924.

In 1927, when the Western Electric Company became associated with the International Telephone and Telegraph Corporation and its name was changed to Standard Telephones and Cables, he became general sales manager and, in 1930, technical director.

He was appointed managing director of Creed & Company in 1933 and served in that capacity until 1957. He became chairman of its board of directors in 1951. At the time of his death, he served also as vice president of International Standard Electric Corporation and as a director of Standard Telephones and Cables, Standard Telecommunication Laboratories, International Marine Radio Company, and Commercial Cable Company, Limited.

During the first world war, he served with distinction in France and in the Near East in King Edward's Horse and the Royal Horse Artillery.

While retaining his active direction of Creed & Company during the second world war, he joined The Ministry of Aircraft Production in 1940 and was principally responsible for safeguarding production through the dispersal of factories and by providing for alternative sources of supply. He became controller of Construction and Regional Services and was a member of the Ministers Council and of the Air Supply Board. As chairman of a joint committee of the three fighting services, he helped to establish methods for equipment preservation and packaging for the Royal Air Force in the tropics. It was for his services to the Royal Air Force during the war that he was made a Knight Bachelor in 1946.

In 1946, he served on the Advisory Panel on Excess Profits Tax Refunds. From 1947 to 1958, he was a director of the British South American Airways Corporation continuing in that capacity when it amalgamated with the British Overseas Airways Corporation. He was appointed to the Board of B.O.A.C. Associated Companies in 1957.

Sir Francis was a Member of the Institution of Electrical Engineers. He held memberships in numerous clubs and was a Liveryman of the City of London, being a member of The Worshipful Company of Gold and Silver Wyre Drawers of the City of London.

He was a life governor of Haileybury and Imperial Service College and a member of the council of Milton Abbey School.

Sir Francis died on June 13, 1960.

Contributors to This Issue

BARRY G. COETSEE

BARRY G. COETSEE was born in 1924. He served in the Royal Navy for two and a half years. He then attended Imperial College of London University and received the B.Sc.,A.C.G.I. degree in 1947.

He joined Standard Telephones and Cables in 1947 as an engineer working on radar display equipment. In 1949, he went to the Canadian General Electric Company in Toronto as a design and production engineer. In 1954, he joined the Radio Corporation of America as an engineer developing transistor mobile communication equipment.

Mr. Coetsee joined ITT Laboratories in 1957 as a senior member of the technical staff that developed the *K24A* carrier system, which is described in this issue.

Mr. Coetsee is a member of both the Association of Professional Engineers of Ontario and of the Institution of Electrical Engineers.

• • •

D. R. COLEMAN was born in Essex, England, in 1925. He received a B.Sc. (Eng.) degree from the University of London; his college attendance was interrupted by service in the Corps of Royal Engineers.

He spent five years on the development of aircraft electrical equipment for the Plessey Company and the Sperry Gyroscope Company.

In 1956, he joined the rectifier division of Standard Telephones and Cables, was in charge of its evaluation and applications section and is now responsible for commercial engineering of rectifier equipments. In this issue, he discusses the application features of silicon junction power rectifiers.

Mr. Coleman is an Associate Member of the Institution of Electrical Engineers and a member of two technical committees for semiconductor rectifiers of the International Electrotechnical Commission.

D. R. COLEMAN

• • •

GEORGE L. CURTIS was born in 1920. He received his B.Sc. degree in electrical engineering from the University of Manitoba in 1940.

From 1940 to 1946, he was with the Bell Telephone Company of Canada; for several years during this period, he was on leave for service with the Royal Canadian Air Force. For 6 years starting in 1946, he was chief engineer of Ontario Northland Communications, a Canadian telephone operating company. In 1952, Mr. Curtis became a project engineer with Lenkurt Electric Company. In 1954, he joined Lynch Carrier Systems as a sales applications engineer.

Mr. Curtis joined ITT Laboratories in 1956; he is commercial product development manager. Some of his work on the *K24A* carrier system is described in this issue.

GEORGE L. CURTIS

Mr. Curtis is a member of the American Institute of Electrical Engineers.

• • •

M. C. DETTMAN was born in New York City on March 5, 1919. He took various courses at New York University, Polytechnic Institute of Brooklyn, College of the City of New York, and RCA Institute.

From 1936 to 1937, Mr. Dettman was employed as supervisory engineer for the National Recording Company. From 1937 to 1942, he was with Radio-marine Corporation of America,

M. C. DETTMAN

SANFORD F. ESSIG

working on all types of marine radio equipment. He was in charge of shipboard installation at Aruba, Dutch West Indies and in Venezuela in 1939.

In 1942, Mr. Dettman joined Federal Telecommunication Laboratories, where he has worked on shipboard and airborne pulse-time-modulation equipment, 500-megacycle-per-second color television, broadcast television transmitting equipment, navar beacon transmitters, microwave pulse-type communication sets, automatic-tuning shipboard communication transmitters, and mobile radio equipment.

In 1958, he transferred to ITT Federal Division as project administrator on the production of the *AN/MRC-55* equipment discussed in this issue. He is currently in administrative engineering.

Mr. Dettman is a member of the Institute of Radio Engineers.

ANATOLE GIRINSKY

SANFORD F. ESSIG received his B.S. degree at Ashland College in 1928 and his M.S. degree at Ohio State University in 1929. After postgraduate work at Ohio State University and several years with Radio Corporation of America, he joined Farnsworth Television and Radio Corporation in 1939.

Mr. Essig has specialized in photosensitive devices, such as pickup tubes, phototubes, multiplier phototubes, image tubes, and infrared devices. He reports in this issue on a high-current-output phototube.

At present, he is head of the phototube group in the ITT Laboratories, Fort Wayne, Indiana.

Mr. Essig is a member of the Optical Society of America, the Electrochemical Society, and the American Institute of Physics.

• • •

ANATOLE GIRINSKY was born on December 27, 1917 in Kinechma, Russia. He graduated as an electrical engineer from Ecole Supérieure d'Electricité of Paris in 1940.

He joined Compagnie Général de Constructions Téléphoniques in 1941, where he has been associated with the development of signaling over power lines. He is now head of the transmission laboratory.

Mr. Girinsky is coauthor of a paper in this issue on high-speed transmission of numerical data over telephone channels.

• • •

JOSEPH W. HALINA was born in 1920. After serving in the Signal Corps of the Royal Canadian Army for five years, he attended the University of Toronto and was awarded his B.Sc. degree in electrical engineering in 1949.

From 1949 to 1953, he was with the General Electric Company doing microwave engineering and later product planning and market research. During 1953–1954, he served the Lenkurt Electric Company as a consulting engineer on microwave problems. For the next two years, he was chief engineer of single-sideband carrier equipment for Lynch Carrier Systems.

In 1956, Mr. Halina joined ITT Laboratories where he had the over-all responsibility for the design of the *K31* rural carrier system and the *K24A* exchange area carrier system, which is described in this issue. He is presently with ITT Communication Systems.

JOSEPH W. HALINA

Mr. Halina is a member of the Association of Professional Engineers of Ontario and of the Institute of Radio Engineers.

• • •

PAUL ROUSSEL was born in Paris, France, on December 23, 1919. He received the Licence es Sciences Physiques from the University of Paris in 1942.

Mr. Roussel joined the technical department of Compagnie Générale de Constructions Téléphoniques in 1943. He is coauthor of a paper on high-speed transmission of numerical data over telephone channels appearing in this issue.

PAUL ROUSSEL

INTERNATIONAL TELEPHONE AND TELEGRAPH CORPORATION

Principal U. S.—Canada Divisions and Subsidiaries

DIVISIONS

Components Division, Clifton, N. J.
Kuthe Laboratories, Inc., Newark, N. J.
Industrial Products Division, San Fernando, Calif.
ITT Distributor Products Division, Lodi, N. J.
ITT Federal Division, Clifton, N. J., and Fort Wayne, Ind.
ITT Kellogg, Chicago, Ill.
ITT Laboratories, Nutley, N. J., and Fort Wayne, Ind.

SUBSIDIARIES

American Cable & Radio Corporation, New York, N. Y.
All America Cables and Radio, Inc., New York, N. Y.
Commercial Cable Company, The, New York, N. Y.
Globe Wireless Ltd., New York, N. Y.
Mackay Radio and Telegraph Company, New York, N. Y.
Federal Electric Corporation, Paramus, N. J.
Northern Services, Inc., Anchorage, Alaska
Intelex Systems Incorporated, New York, N. Y.
Airmatic Systems Corporation, Rochelle Park, N. J.
International Electric Corporation, Paramus, N. J.
ITT Communication Systems, Inc., Paramus, N. J.
IT&T Electronics Service Company of Canada Ltd., Town of Mount Royal, P.Q.
Kellogg Credit Corporation, New York, N. Y.
Royal Electric Corporation, Pawtucket, R. I.
Standard Telephones & Cables Mfg. Co. (Canada) Ltd., Montreal, P. Q.

Principal overseas research, manufacturing, and sales affiliates

ARGENTINA
Capehart Argentina S.A.I.C. (50% owned), Buenos Aires
Compañia Standard Electric Argentina, S.A.I.C., Buenos Aires

AUSTRALIA
Standard Telephones and Cables Pty. Limited, Sydney
Austral Standard Cables Pty. Limited (50% owned), Melbourne

AUSTRIA
Standard Telephon und Telegraphen Aktiengesellschaft, Czeija, Nissl & Co., Vienna

BELGIUM
Bell Telephone Manufacturing Company, Antwerp
ITT Europe Inc., Brussels

BRAZIL
Standard Eléctrica, S.A., Rio de Janeiro

CHILE
Compañía Standard Electric, S.A.C., Santiago

CUBA
Equipos Telefónicos Standard de Cuba, Havana

DENMARK
Standard Electric Aktieselskab, Copenhagen

FINLAND
Oy Suomen Standard Electric AB, Helsinki

FRANCE
Compagnie Générale de Constructions Téléphoniques, Paris
Les Téléimprimeurs, Paris
International Standard Engineering, Inc., Paris
Laboratoire Central de Télécommunications, Paris
Le Matériel Téléphonique, Paris

GERMANY
Standard Elektrik Lorenz Aktiengesellschaft, Stuttgart
Divisions:
Bauelemente Werk S.A.F., Nuremberg
Informatikwerk, Stuttgart
Kabelwerk, Stuttgart
Lorenz Werke, Stuttgart
Mix & Genest Werke, Stuttgart
Schaub Werk, Pforzheim

IRAN
Standard Electric Iran A.G., Teheran

ITALY
Fabbrica Apparecchiature per Comunicazioni Elettriche Standard S.p.A., Milan

MEXICO
Industria de Telecomunicación, S.A. de C.V. (50% owned), Mexico City
Standard Eléctrica de México, S.A., Mexico City

NETHERLANDS
Nederlandsche Standard Electric Maatschappij N.V., The Hague

NORWAY
Standard Telefon og Kabelfabrik A/S, Oslo

PORTUGAL
Standard Eléctrica, S.A.R.L., Lisbon

SPAIN
Standard Eléctrica, S.A., Madrid

SWEDEN
Standard Radio & Telefon AB, Stockholm

SWITZERLAND
Standard Téléphone et Radio S.A., Zurich

TURKEY
Standard Elektrik ve Telekomünikasyon Limited Şirketi, Ankara

UNION OF SOUTH AFRICA
Standard Telephones and Cables (South Africa) (Proprietary) Limited, Boksburg, Transvaal

UNITED KINGDOM
Creed & Company Limited, Croydon
Standard Telephones and Cables Limited, London
Kolster-Brandes Limited, Sidcup
Standard Telecommunication Laboratories Limited, London

VENEZUELA
Standard Telecommunications C.A., Caracas

OVERSEAS TELECOMMUNICATION COMPANIES

ARGENTINA
Compañía Internacional de Radio, S.A., Buenos Aires
Sociedad Anónima Radio Argentina (subsidiary of American Cable & Radio Corporation), Buenos Aires

BOLIVIA
Compañía Internacional de Radio Boliviana, La Paz

BRAZIL
Companhia Rádio Internacional do Brasil, Rio de Janeiro
Companhia Telefônica Nacional, Curitiba and Pôrto Alegre

CHILE
Compañía de Teléfonos de Chile, Santiago
Compañía Internacional de Radio, S.A., Santiago

CUBA
Cuban American Telephone and Telegraph Company (50% owned), Havana
Cuban Telephone Company, Havana
Radio Corporation of Cuba, Havana

PERU
Compañía Peruana de Teléfonos Limitada, Lima

PUERTO RICO
Puerto Rico Telephone Company, San Juan
Radio Corporation of Puerto Rico, San Juan

SPAIN
Compañía Radio Aérea Marítima Española, S.A., Madrid

UNITED KINGDOM
International Marine Radio Company Limited, Croydon

VIRGIN ISLANDS
Virgin Islands Telephone Corporation, Charlotte Amalie

ASSOCIATE LICENSEES FOR MANUFACTURE AND SALES

FRANCE
Lignes Télégraphiques et Téléphoniques, Paris

ITALY
Società Italiana Reti Telefoniche Interurbane, Milan

JAPAN
Nippon Electric Company, Limited, Tokyo
Sumitomo Electric Industries, Limited, Osaka

SPAIN
Marconi Española, S.A., Madrid

IN THIS ISSUE

Modern Power–Tube Techniques

K24A Syncroplex Telephone
Carrier System

High-Speed Data Transmission
Over Telephone Channels

Performance of Silicon Junction
Power Rectifiers

Phototubes Capable of High Current Output

AN/MRC-55 Mobile High-Frequency
Communication Equipment

VOLUME 36 1960 NUMBER 4

ELECTRICAL COMMUNICATION

INTERNATIONAL TELEPHONE AND TELEGRAPH CORPORATION

Printed in the
United States of America
LANCASTER PRESS, INCORPORATED
LANCASTER, PENNSYLVANIA